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The Proceedings

OF

THE INSTITUTION OF ELECTRICAL ENGINEERS

FOUNDED 1871: INCORPORATED BY ROYAL CHARTER 1921

PART B

ELECTRONIC AND COMMUNICATION ENGINEERING
(INCLUDING RADIO ENGINEERING)

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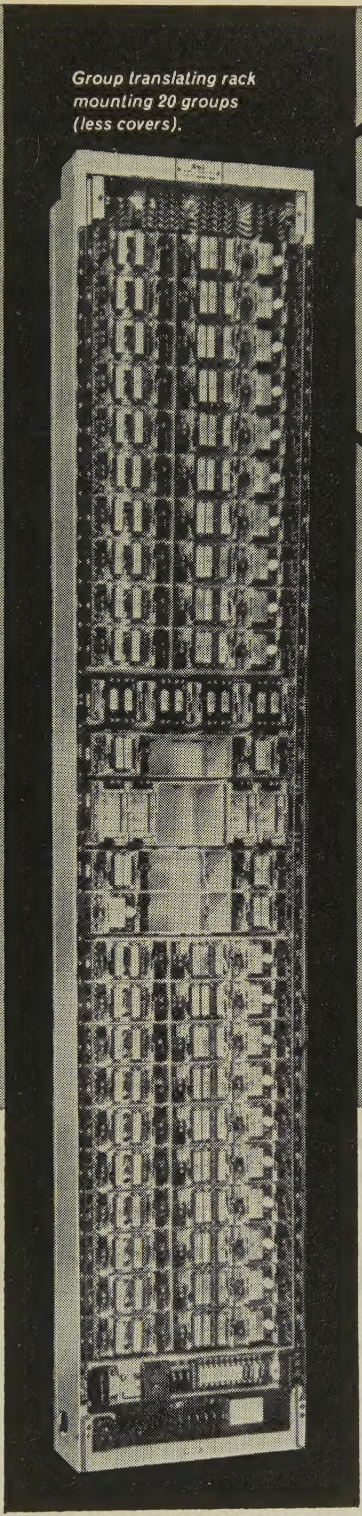

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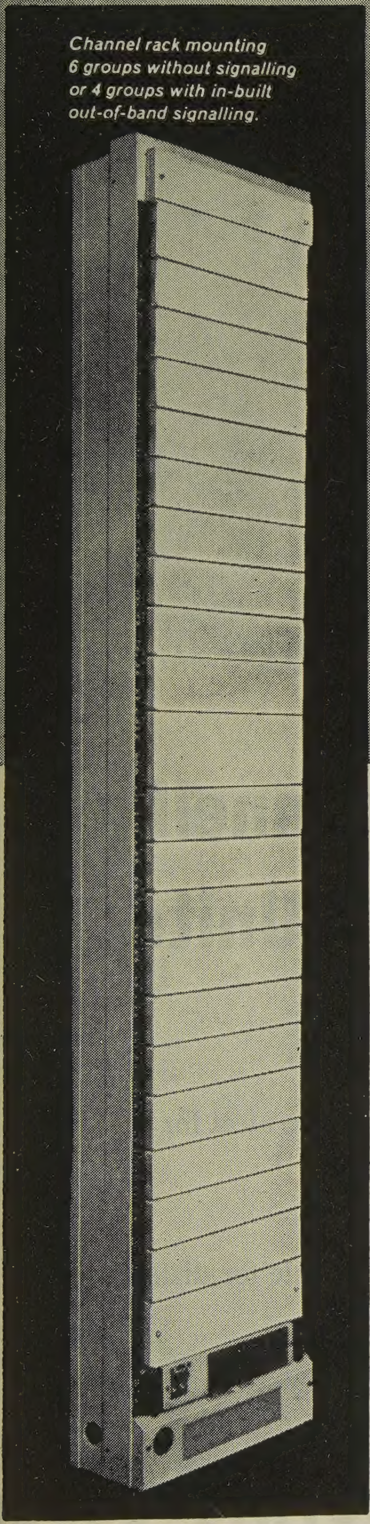

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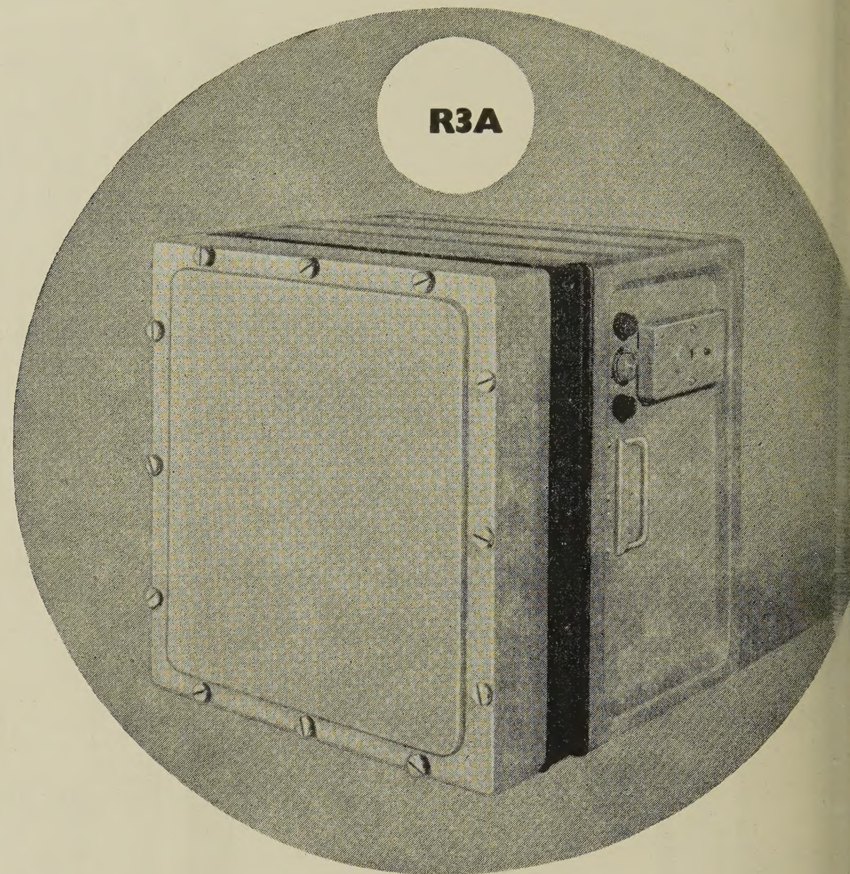

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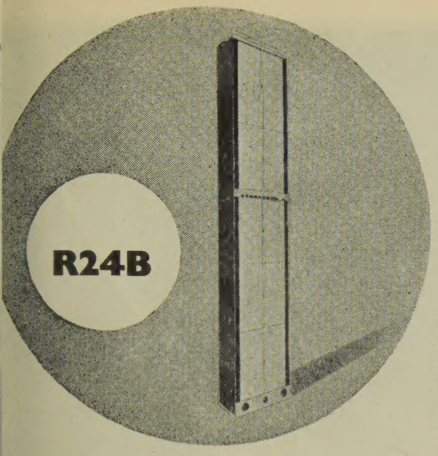

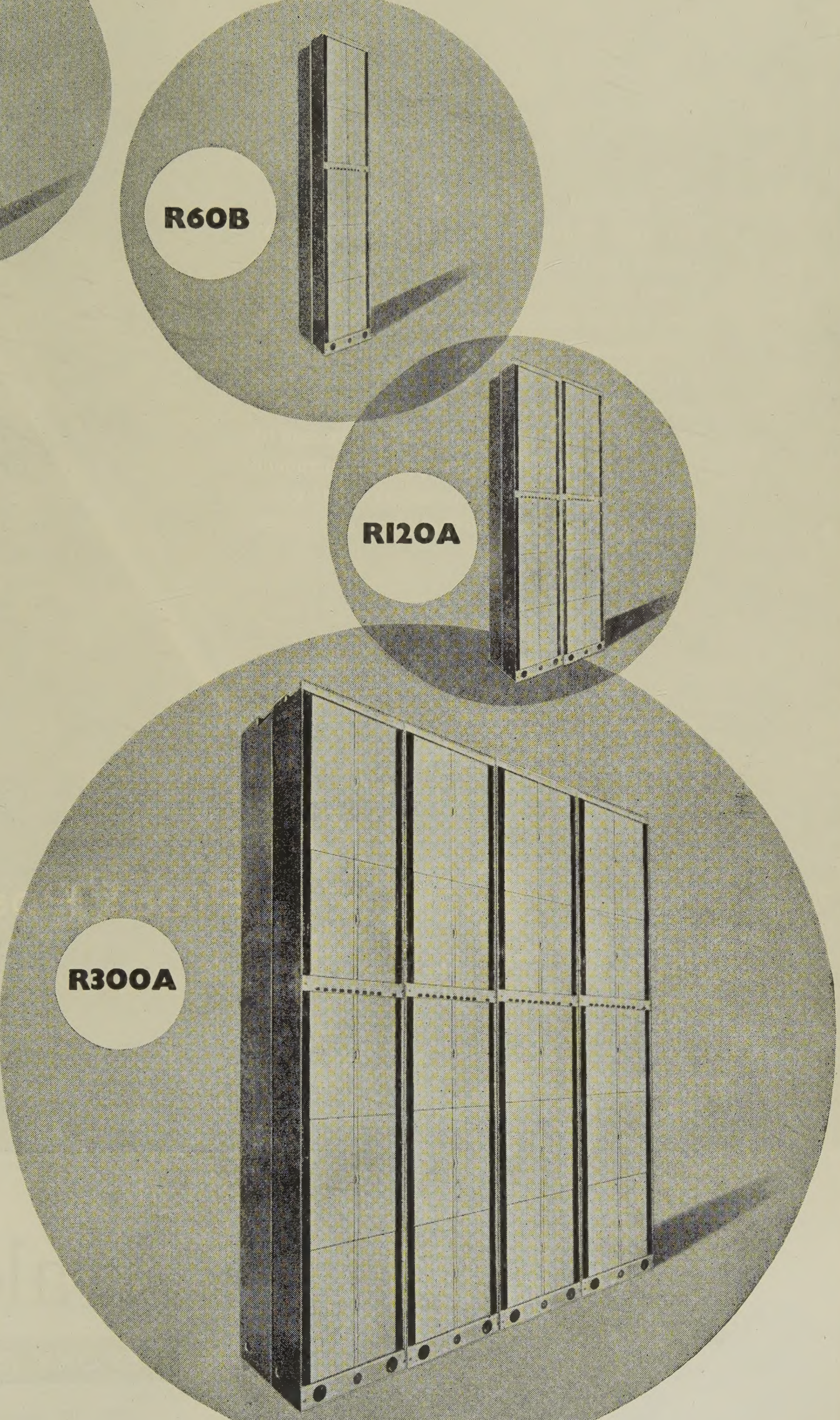

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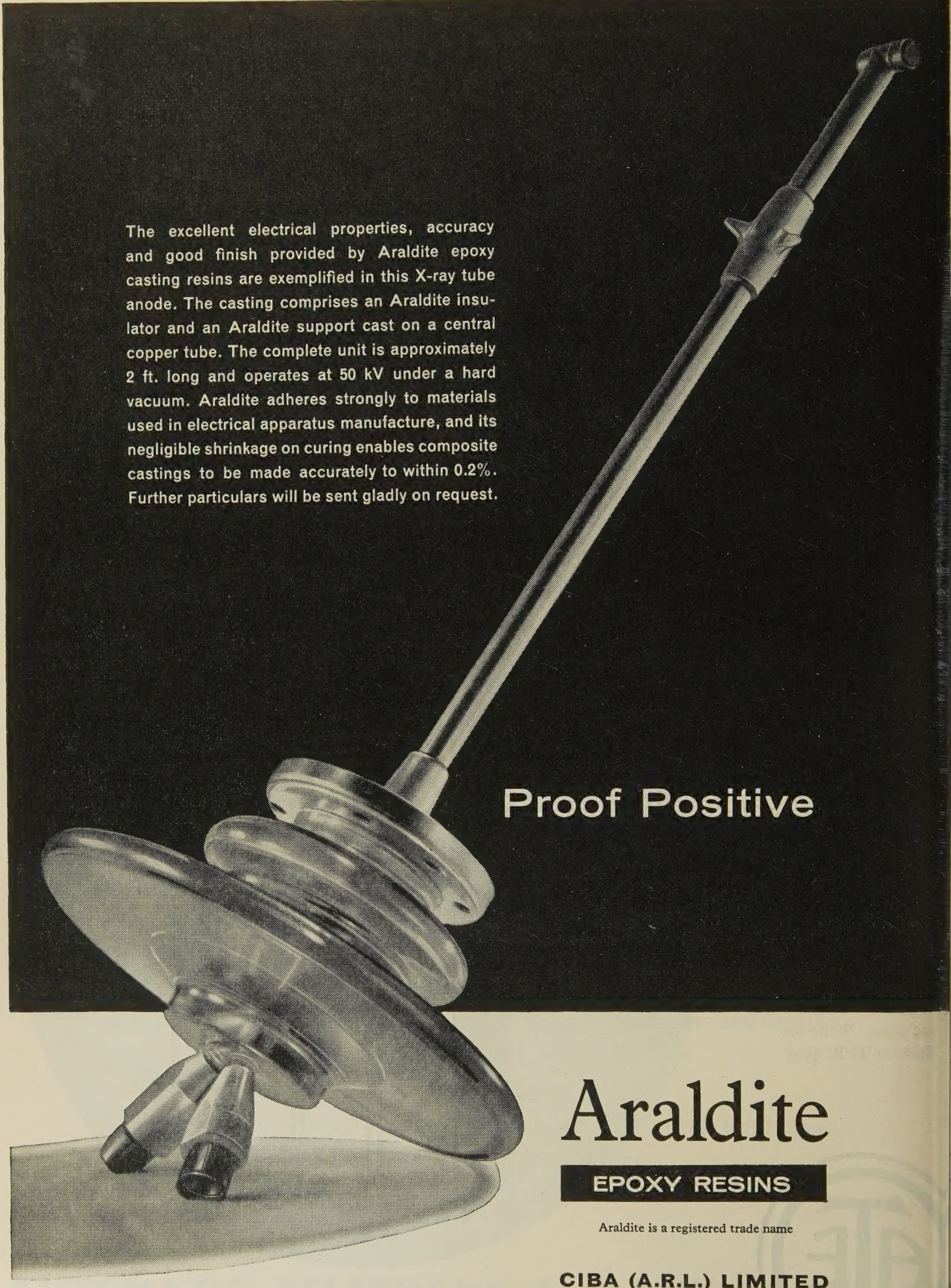

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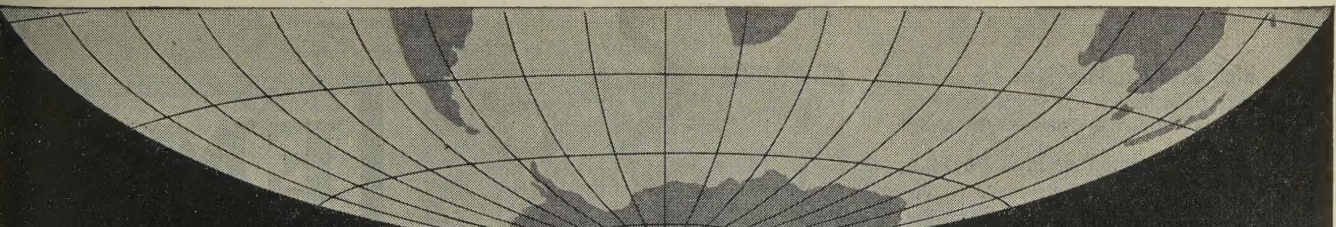

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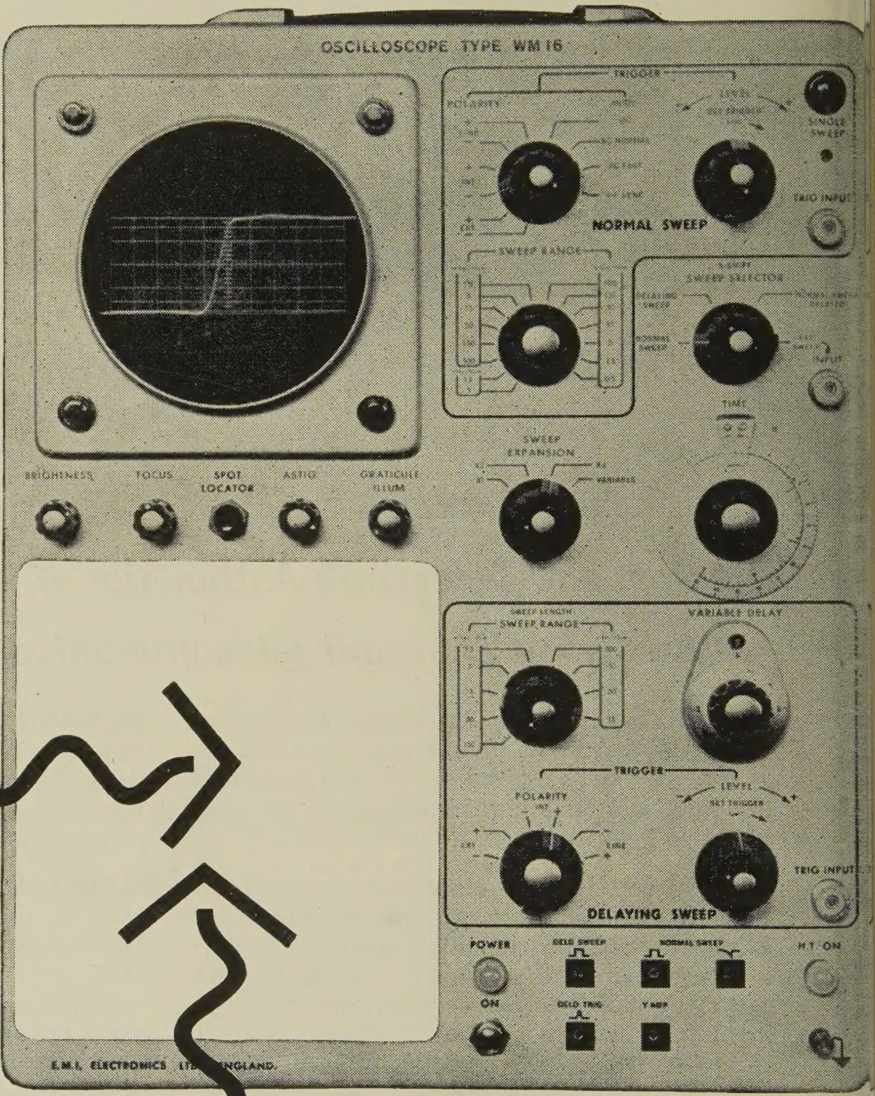

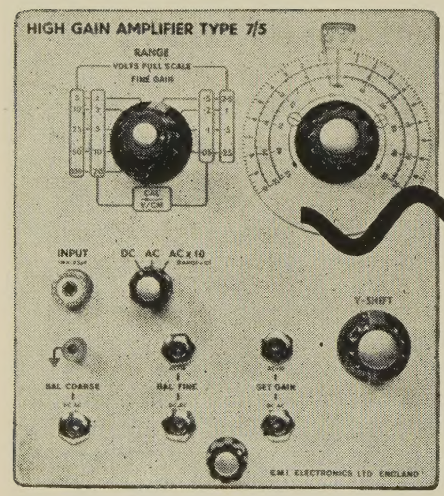

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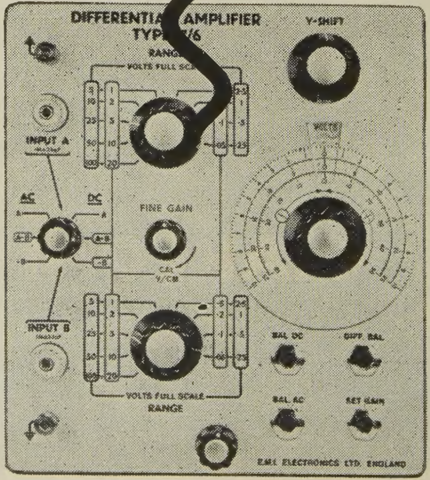

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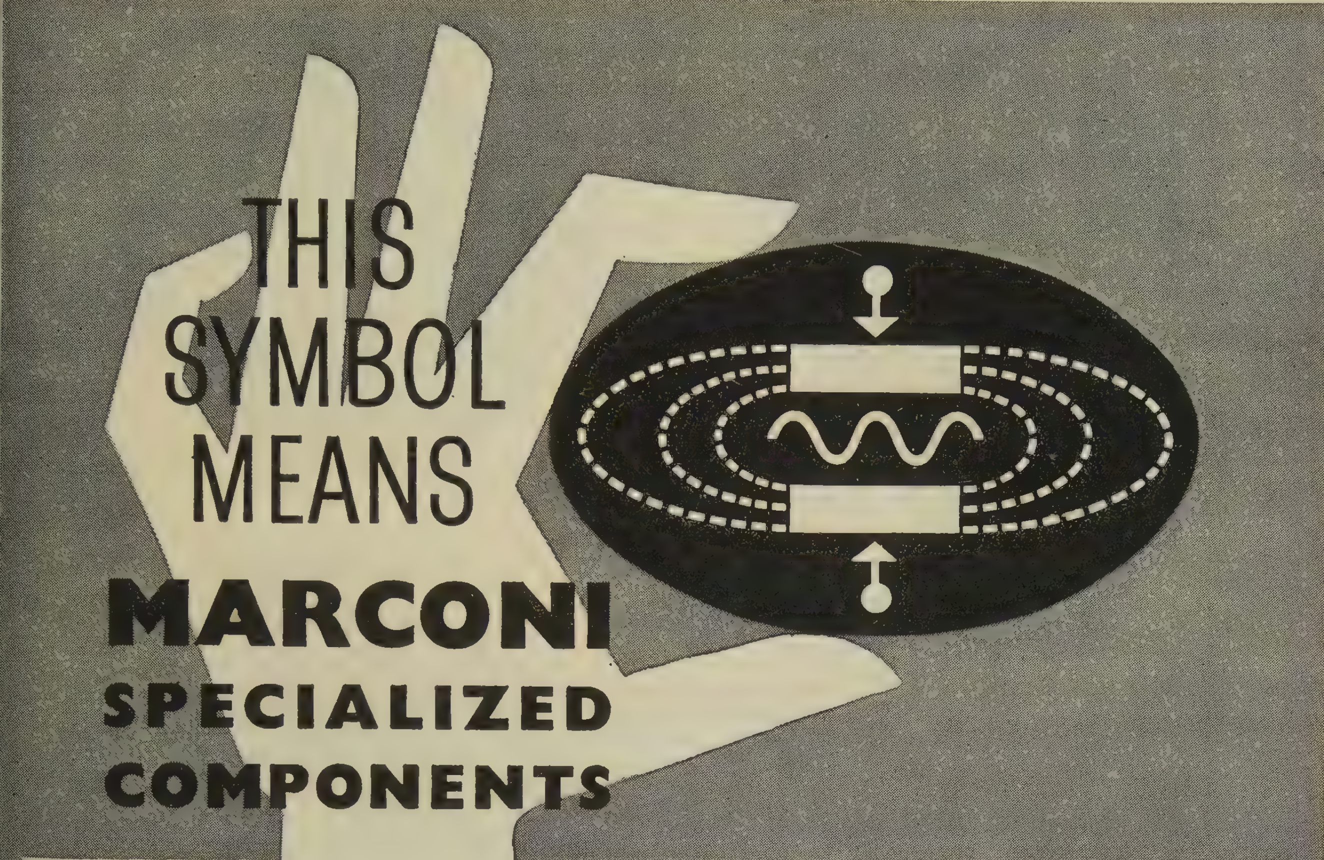

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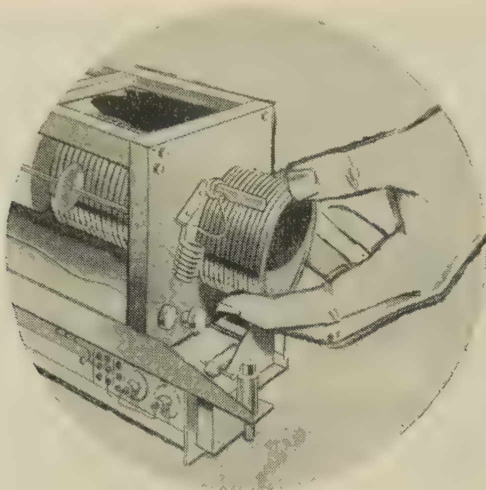

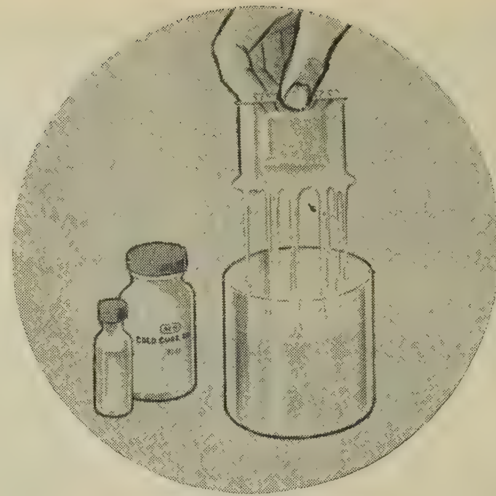

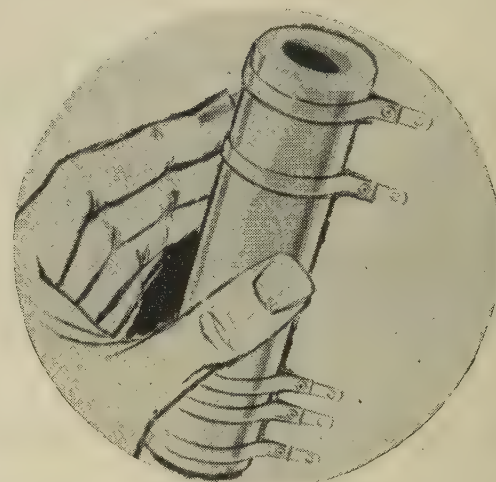

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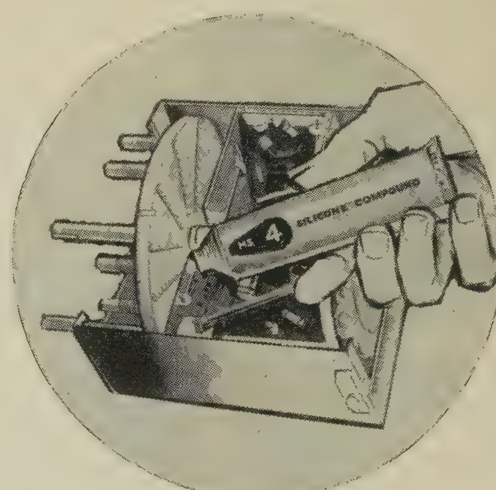

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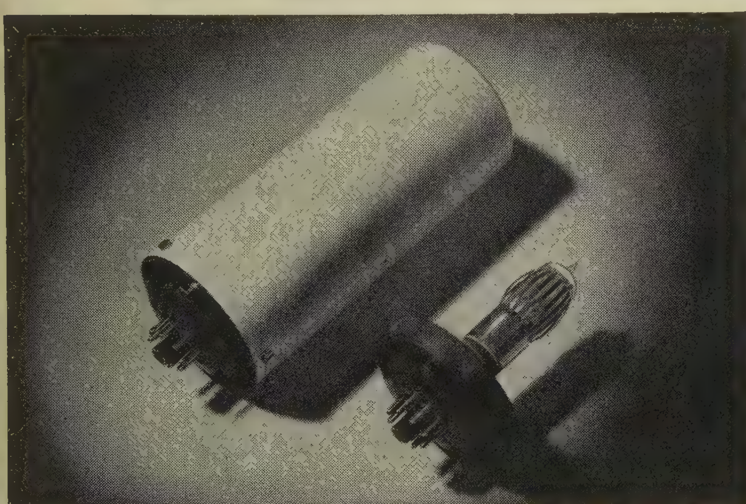

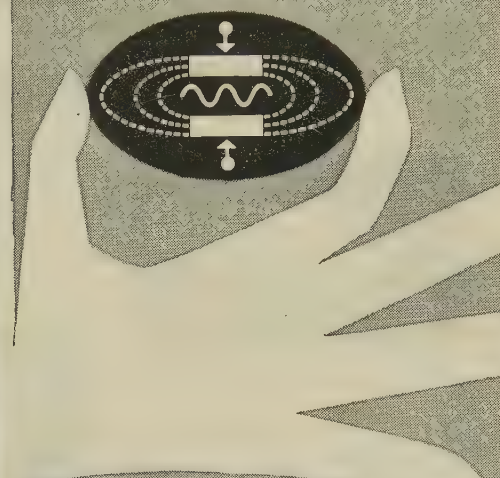

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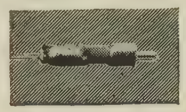
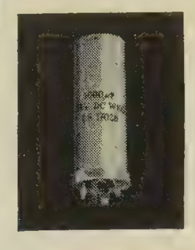
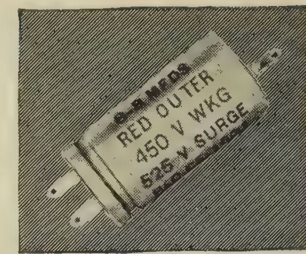
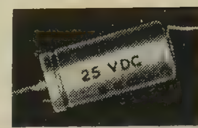
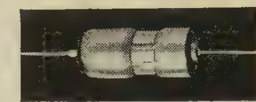
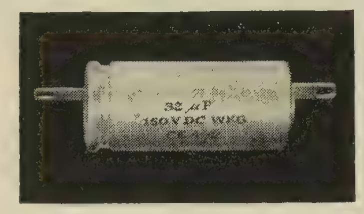
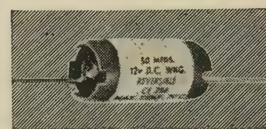
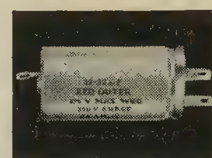




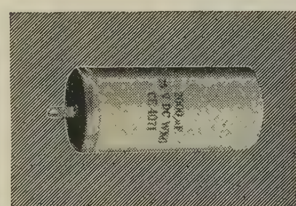
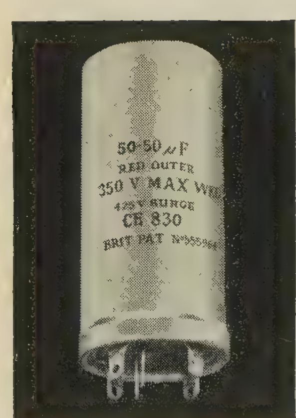




Capacitorability



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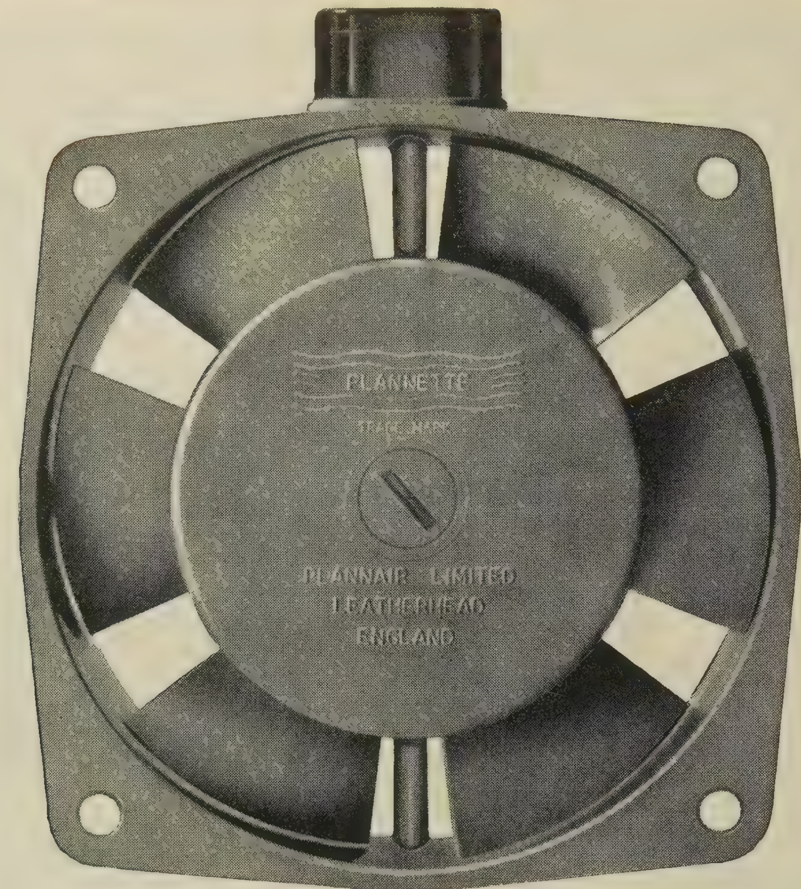

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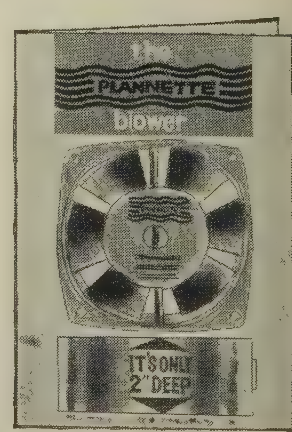

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2,700 r.p.m.:	1,400 r.p.m.:
4½″ delivers 80 c.f.m. at .15 s.w.g.	4½″ delivers 40 c.f.m. at .04 s.w.g.
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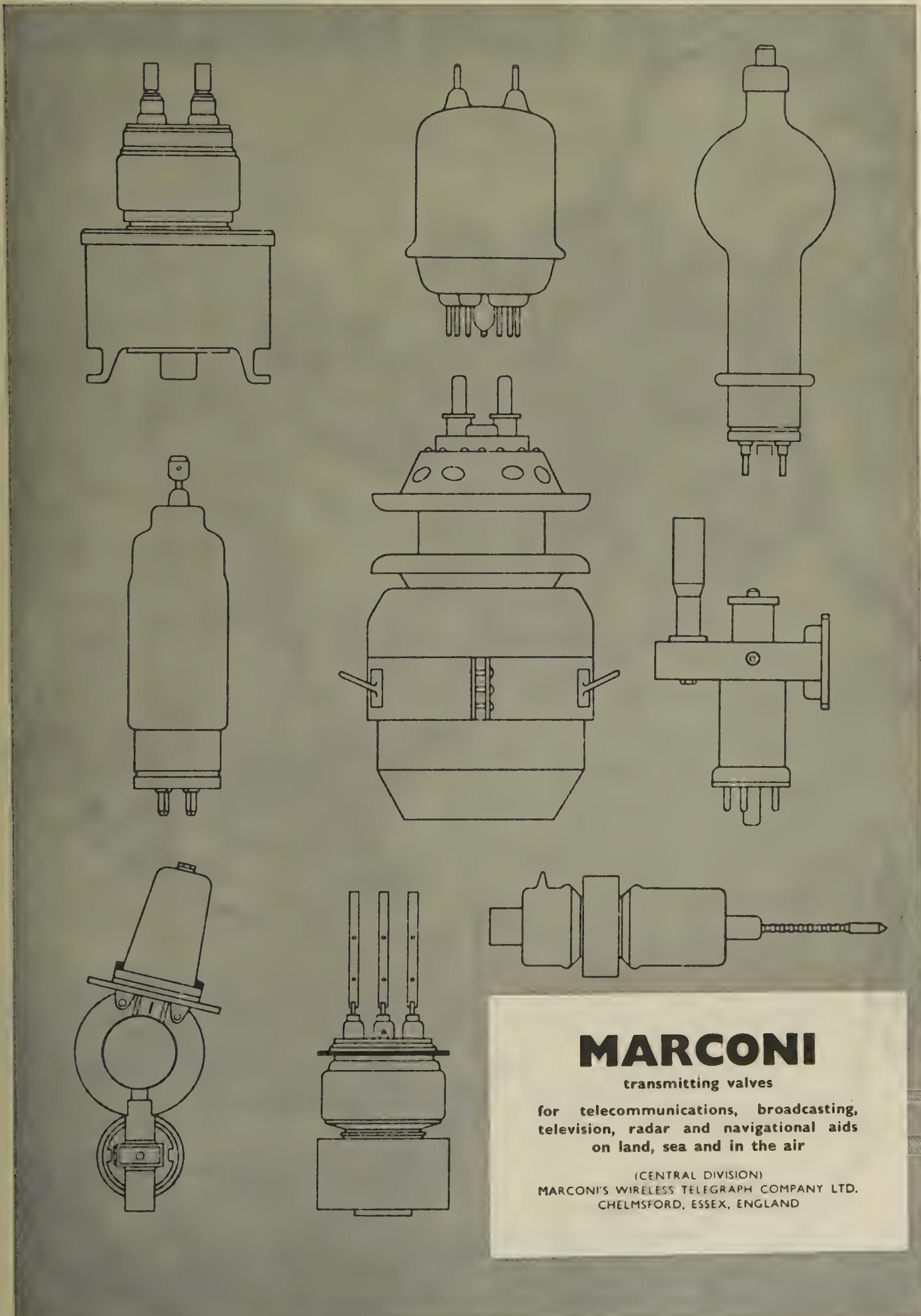

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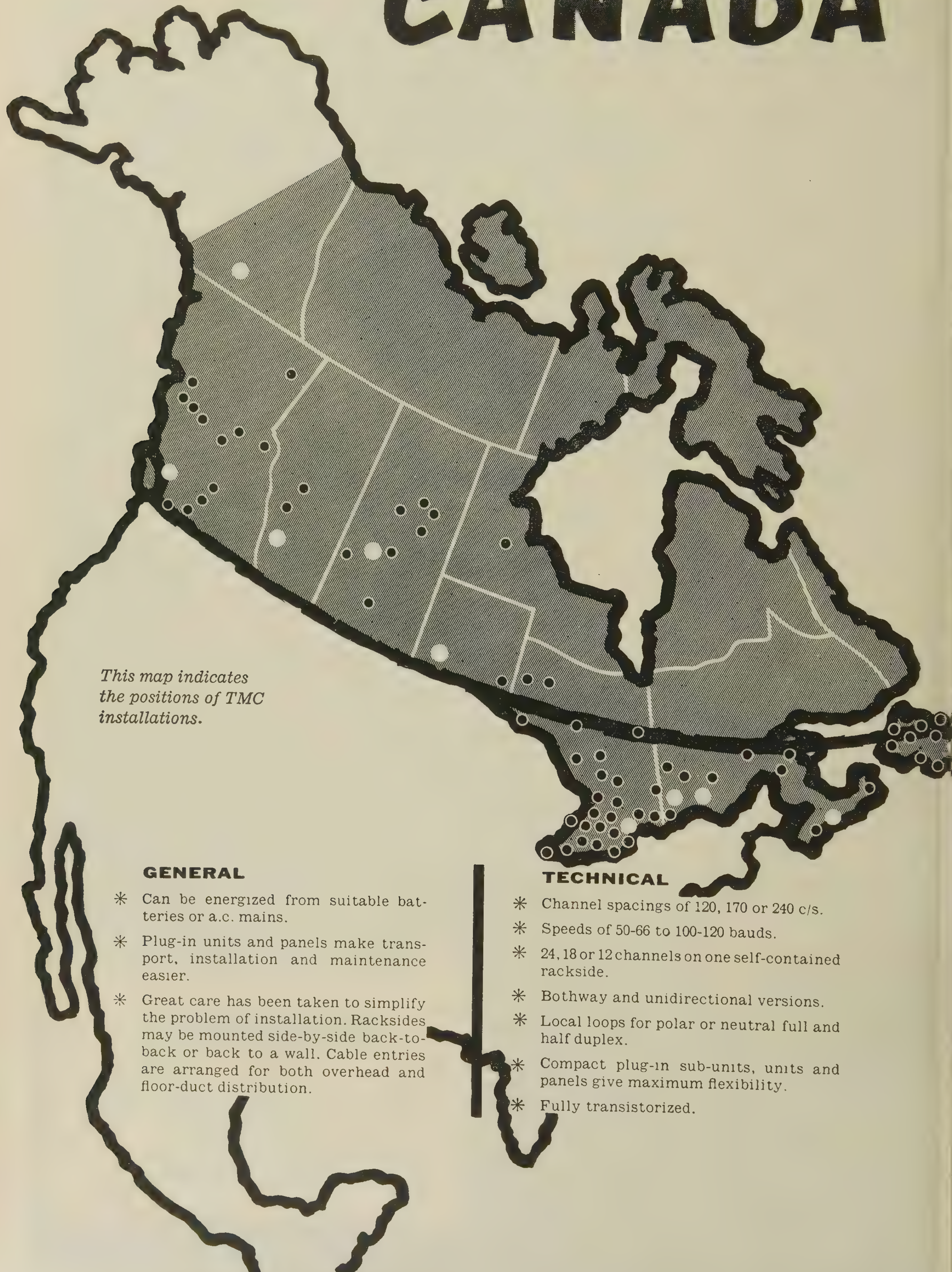

This map indicates the positions of TMC installations.

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* Can be energized from suitable batteries or a.c. mains.
* Plug-in units and panels make transport, installation and maintenance easier.
* Great care has been taken to simplify the problem of installation. Racksides may be mounted side-by-side back-to-back or back to a wall. Cable entries are arranged for both overhead and floor-duct distribution.

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* Channel spacings of 120, 170 or 240 c/s.
* Speeds of 50-66 to 100-120 bauds.
* 24, 18 or 12 channels on one self-contained rackside.
* Bothway and unidirectional versions.
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* Compact plug-in sub-units, units and panels give maximum flexibility.
* Fully transistorized.

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TMC transistorized frequency modulated V.F. Telegraph Equipment is an acknowledged contribution to the recent advances made by the electronics industry in the field of telecommunications. In comparison with earlier designs, it effects considerable economies by using less space, reducing the power consumed and being simple to install as well as easy to maintain.

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TMC

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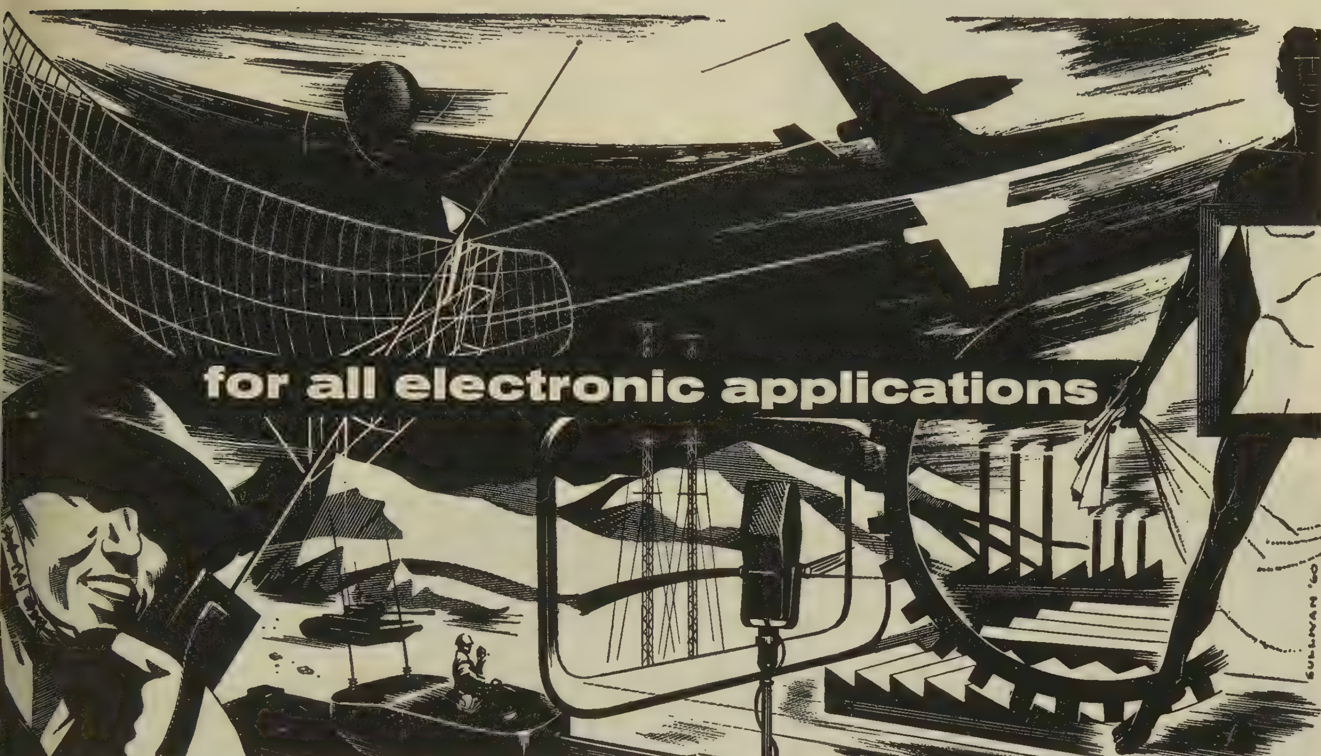

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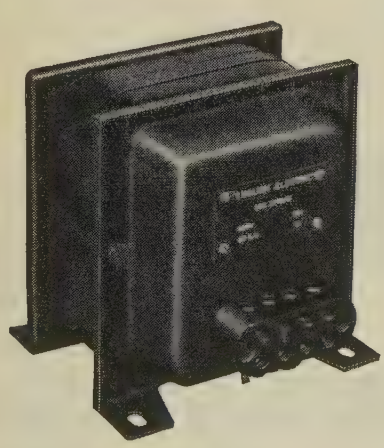

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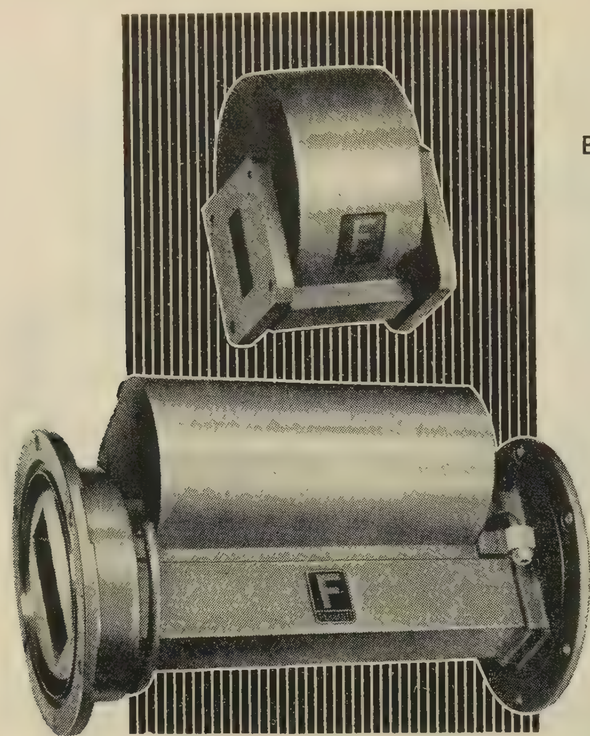

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Bandwidth about the centre frequency	±5%
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Isolation at centre frequency	35db
Isolation over the band	>15db
Weight	2½lb approx

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Mean Power	3·5 KW
Isolation at centre frequency	Not less than 12d
Isolation over the band	Not less than 9d
Weight	30lb

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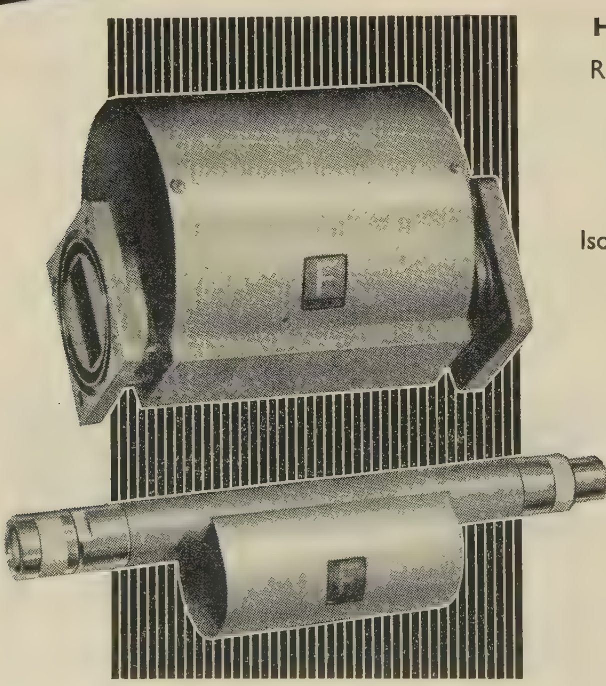

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Range of centre frequencies	8500-10,500 Mc/s
Bandwidth	±5% about centre frequen
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Mean Power	250 W (pressurised)
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Range of centre frequencies	2500-4500 Mc/s
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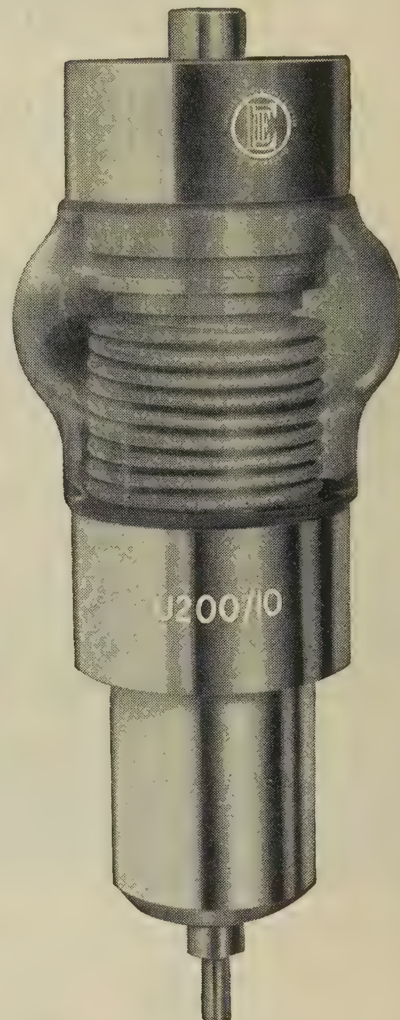

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* **Easily demountable.**

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U200/8	5.5—206	17	8	20†	9.5	2.49
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* Up to 30 Mc/s † Up to 20 Mc/s

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Voltage Range: 6 to 100V. d.c.
Capacitance Range: 0.2 to 200 µF.

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TANTALUM ELECTROLYTIC CAPACITORS

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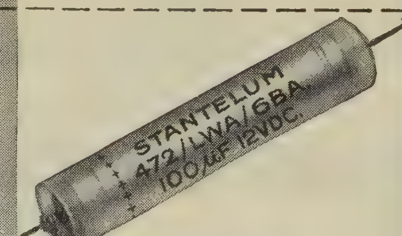
Type approved to R.C.S.134B
Temperature Range: -40 to +85°C.
Voltage Range: 6 to 150V. d.c.
Capacitance Range: 0.15 to 200 µF.



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(POLAR)

Sintered Slug and solid electrolyte construction.
Temperature Range: -55 to +85°C.
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Capacitance Range: 1 to 220 µF.

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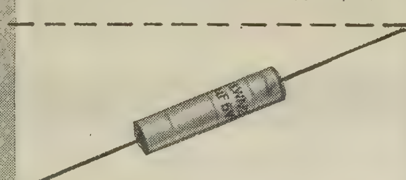
A foil type tantalum capacitor in its most economical form.
Temperature Range: -25 to +70°C.
Voltage Range: 3 to 25V. d.c.
Capacitance Range: 1.5 to 16 µF.



STC manufacture a full range of tantalum electrolytic capacitors.

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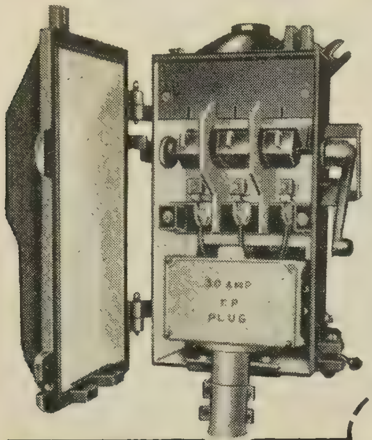

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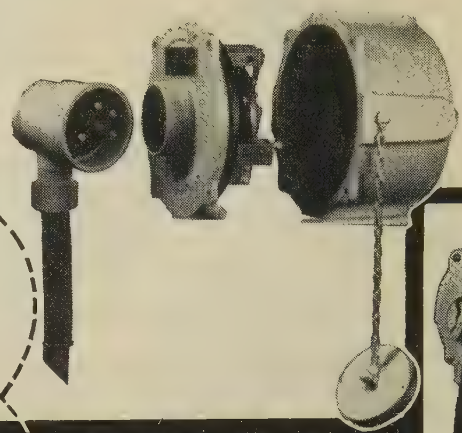

30 to 100 amp. 500 volt A.C. or D.C., two or three-pole splash-proof switch interlocked with lid. (Form B).

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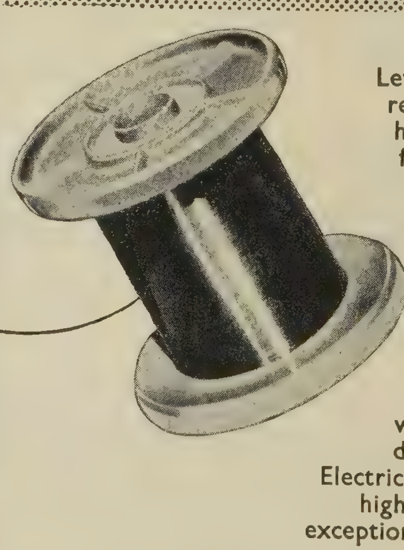

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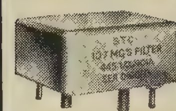

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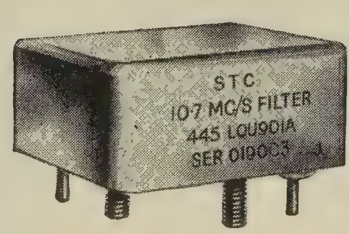

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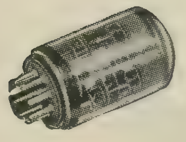

Illustrating the potting of silicon junction diodes, the diodes being assembled on to a standard international octal valve base thus providing a plug-in rectifier capable of .25 A.D.C. current with a P.I.V 800-800.

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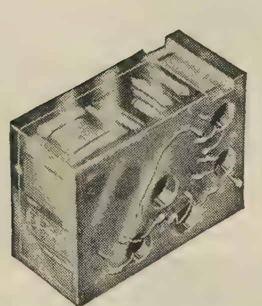

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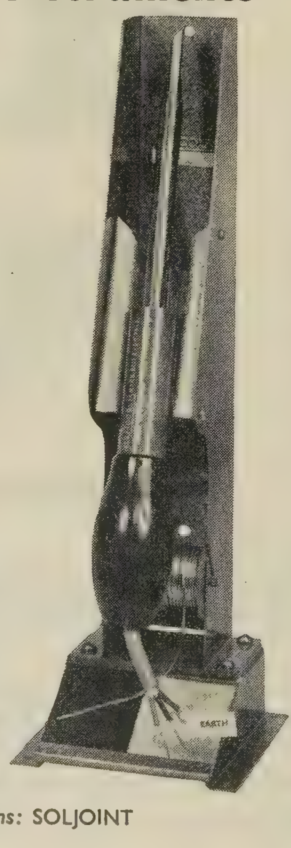

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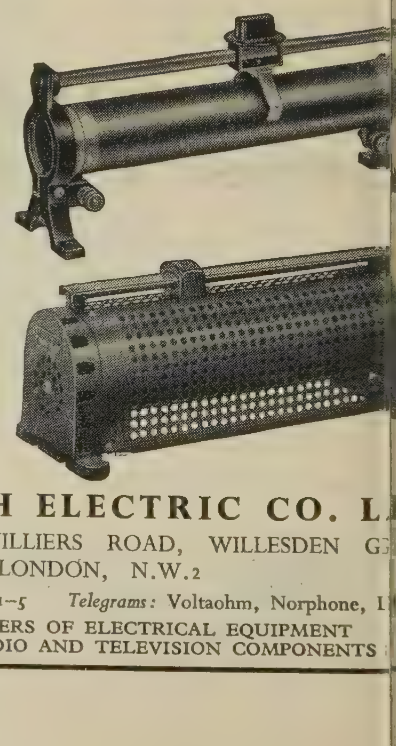

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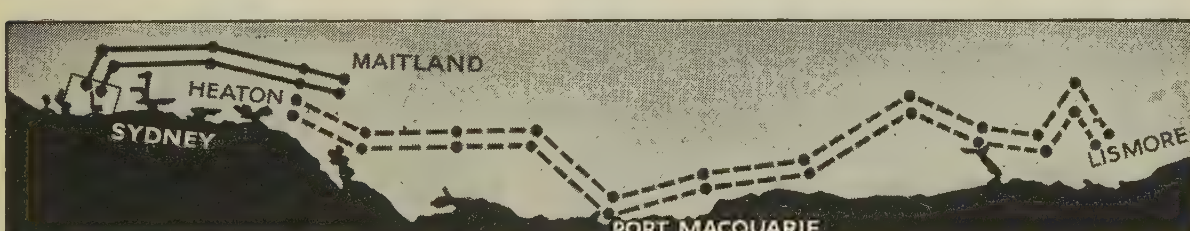

Fully duplicated radio channels will be equipped with automatic baseband switching at the terminals. The Heaton and Lismore system will employ periodic magnet travelling wave amplifiers, and will also include the new STC Type RA4 auxiliary link providing remote indication and control circuits over a separate microwave channel.

STC are supplying main line microwave telephone systems to 18 countries and have already supplied systems with a capacity of nearly $5\frac{1}{2}$ million telephone circuit miles and nearly 6 000 television channel miles.

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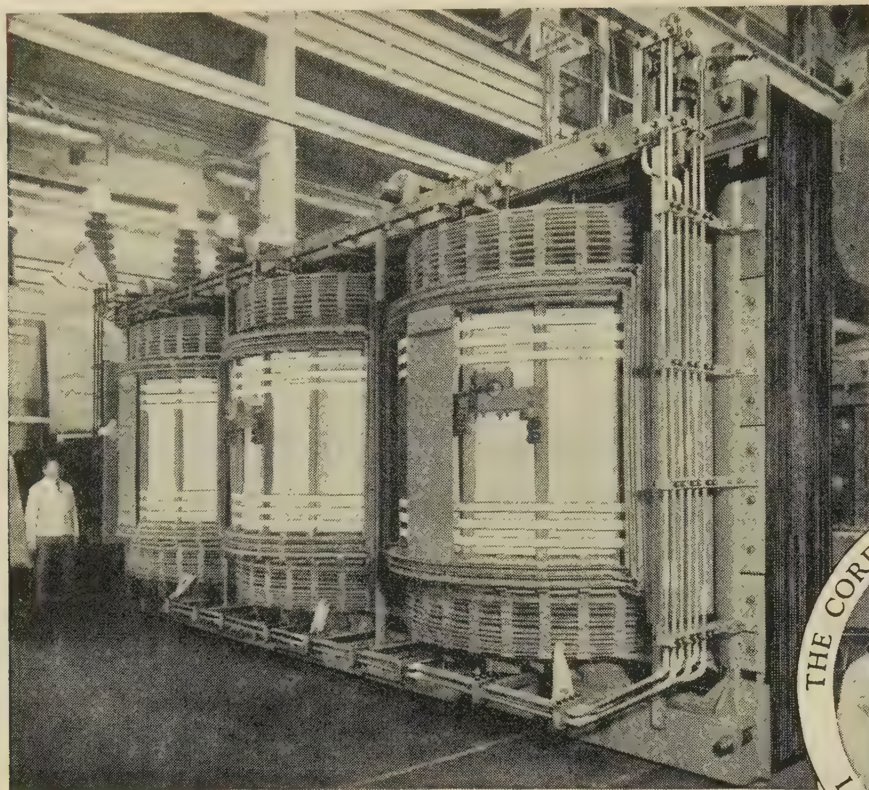

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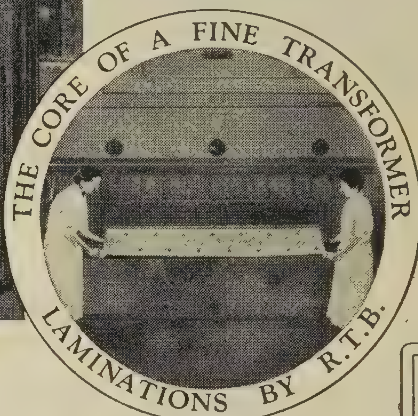

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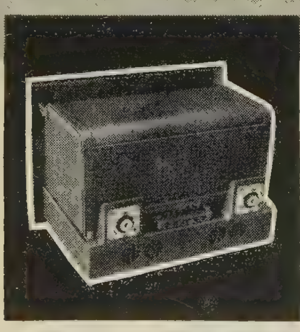

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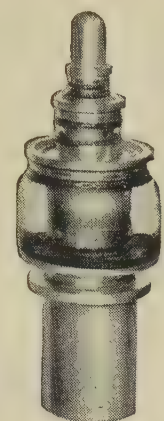
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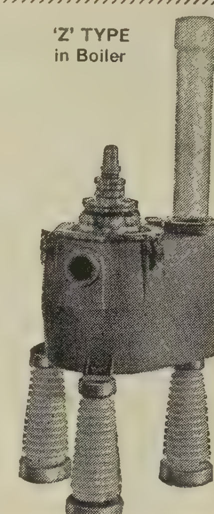


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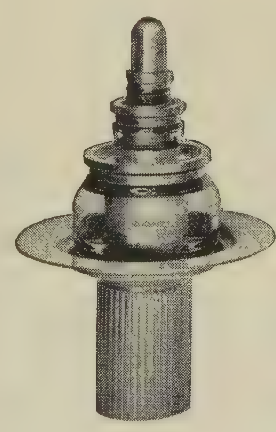
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Proceedings of the First International Congress of the International Federation of Automatic Control (IFAC)—MOSCOW 1960

Editor: J. F. COALES · *Co-editors:* J. R. RAGAZZINI *and* A. T. FULLER

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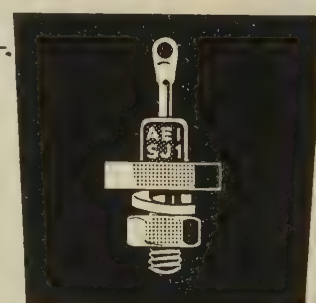

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The Institution of Electrical Engineers
Paper No. 3508 E
May 1961

STUDIES OF IONOSPHERIC FORWARD SCATTERING USING MEASUREMENTS OF ENERGY DISTRIBUTION IN AZIMUTH

By W. C. BAIN, M.A., B.Sc., Ph.D., Associate Member.

(*The paper was first received* 7*th January, in revised form* 20*th July, and in final form* 28*th November*, 1960.)

SUMMARY

Measurements of phase and amplitude have been carried out on nals on 37 Mc/s received from a transmitter at Gibraltar. The eiving sites were at Slough and Castlemartin, which are each about 40 km from Gibraltar, and were equipped with pairs of aerials ose spacing could be varied from λ to 10λ. The results were not sistent with the idea that the forward-scattered signal was due to combination of many randomly-phased radio waves. Frequently signal appeared to be dominant, as might be expected if reflections m meteor trails formed an important part of the resultant signal. calculation of an azimuth power distribution on the basis of phase asurements with different aerial spacings then becomes of uncertain idity. Some results obtained in this way are presented, but reliance placed mainly on work with a small, fixed aerial spacing. This ws a marked diurnal variation of mean bearing; at Slough for st of the year it is on average 7°W of the Gibraltar bearing by night 0°W by day. A comparison with observations at Castlemartin in les suggests that the scattering process is due almost entirely to eor reflections by night but that during the day there is also present mponent due to turbulent scattering which contributes rather more n half the total energy.

LIST OF PRINCIPAL SYMBOLS

y = Aerial spacing in wavelengths.
α = Bearing relative to great-circle direction of Gibraltar.
$s = \sin\alpha$.
ψ_k = The kth measured phase difference.
ϕ_k = The kth phase difference (from a sample with zero mean).
$\phi)$ = Theoretical distribution of phase differences ϕ.
ν = A mean value of phase differences ψ_k [see eqn. (1)].
R = A measure of the spread of phase differences ψ_k [see eqn. (2)].
ρ_A = Amplitude correlation coefficient between signals at the two aerials.
σ = Standard deviation of a Gaussian angular power spectrum.
p_k, q_k = Lengths of major and minor axes of the kth ellipse on the phase display.
$\beta = \cos^{-1} R$.
X = Scattering angle.

ritten contributions on papers published without being read at meetings are ed for consideration with a view to publication.
e paper is an official communication from the Radio Research Station, Depart- of Scientific and Industrial Research.

(1) INTRODUCTION

In a previous paper[1] some preliminary results were given concerning the distribution of energy in azimuth which is created by the ionospheric scattering process at very high frequencies. This paper contains a description of the whole of the work carried out on this project, and of the results obtained from it; some of the details given in the earlier paper will be repeated here for convenience.

The object of the experiments was to measure at various times of day and of the year the angular power spectrum of the incident radiation, as the distribution of energy in azimuth is often called. This should enable one to calculate the likely performance of aerials to be used on ionospheric forward-scatter circuits. It was also hoped that the azimuth distributions obtained would throw some light on the nature of the scattering regions, for there has been some doubt in the past about the extent to which these are due to meteors or to turbulence in the ionosphere.

The term 'forward-scatter' in this paper will be used to describe the signal which is usually received at distances of 1 000–2 000 km from a v.h.f. transmitter, excluding only those times when a single large meteor signal is present or when propagation is by way of the normal ionospheric layers or the sporadic-E layer. It is not intended to imply by the use of this term that turbulent scattering is necessarily the operative process.

Bearings in the following Sections will be quoted relative to the great-circle direction of Gibraltar, except where otherwise stated. All times are Universal Time (U.T.).

(2) THEORY OF THE METHOD

The received signals were picked up on a pair of aerials situated in a line normal to the great-circle path from the transmitter and measurements were made in rapid succession of the phase difference between the signal outputs from the two aerials. If the incident radiation contains waves arriving simultaneously from many directions in space and randomly related

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in phase, the desired angular power spectrum can be calculated from these measurements by an extension of some results obtained by Bramley.[2] The method requires a knowledge of the distributions of phase differences which would be obtained for all aerial spacings from zero to infinity; of course, in practice, a finite number of spacings has to be used. Each of these distributions is theoretically of a particular mathematical form, as in eqn. (8), and is specified completely by the parameters ν and R which are defined below. They have to be determined at each aerial spacing from the observations. As shown in Section 11.1, the angular power spectrum can then be calculated from the values of ν and R at the aerial spacings used.

Suppose that n measurements of phase difference, ψ_k, are made at a particular aerial spacing of y wavelengths. ν and R are defined as

$$\nu = \arg\left(\sum_{k=1}^{n} \exp j\psi_k\right) \quad . \quad . \quad . \quad . \quad . \quad (1)$$

$$R = \cos\left(\frac{1}{n}\sum_{k=1}^{n} |\psi_k - \nu + 2m\pi|\right) \quad . \quad . \quad . \quad (2)$$

where $m = 1, 0, -1$ as required to make $|\psi_k - \nu + 2m\pi| < \pi$. ν can be considered to be a kind of mean value for the phase differences, whereas R is a measure of their spread. R is unity where the spread of the phase differences is negligibly small, and is zero when the spread is uniform over 2π radians.

A number of different aerial spacings are used, so that ν and R are functions of y. If $s = \sin\alpha$, where α is the azimuth measured from the direction of the transmitter, the angular power spectrum Q(s) as a function of sin α is given by

$$Q(s) = 2\int_0^\infty (R\cos\nu\cos 2\pi ys + R\sin\nu\sin 2\pi ys)dy \quad . \quad (3)$$

If Q(s) is symmetrical about a particular value, s_0, then as is shown in Section 11.2, ν is related to s_0 by

$$\nu = 2\pi y s_0 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

Also, if no energy is arriving from angles more than 30° off the great-circle path, then, to within 1°, the mean bearing of the incident radiation is given by

$$\bar{\alpha} = \sin^{-1} s_0 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

The theory is given in more detail in Section 11.1.

Now suppose that ν is obtained on a particular occasion with a small aerial spacing. For any angular power spectrum, Q(s), a value of s_0 can be obtained by using eqn. (4) and this can be taken to be a mean value for s (= sin α)—not, in general, the true arithmetic mean unless Q(s) is symmetrical, but one which approaches it as the aerial spacing is reduced. As s is very nearly equal to α for all bearings from which energy is coming, the value of $\bar{\alpha}$ given by eqn. (5) can be considered to give a mean bearing and will frequently be employed in this sense in the following Sections, generally when an aerial spacing of λ has been used.

Certain other results given by Bramley[2] will be required here. The amplitude correlation coefficient ρ_A, which is that between the envelopes of the voltages at the two aerials, is related to the function R by the formula

$$\rho_A = 0\cdot91\left(R^2 + \frac{R^4}{16} + \ldots\right) \quad . \quad . \quad . \quad . \quad (6)$$

The distribution of phase differences is also useful. If

$$\phi_k = \psi_k - \nu \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

then the distribution of ϕ, from which each ϕ_k may b[e con]sidered to be drawn, is given by

$$p(\phi) = \frac{1 - R^2}{2\pi}\left\{\frac{1}{1 - R^2\cos^2\phi} + \frac{R\cos\phi}{(1 - R^2\cos^2\phi)^{3/2}}\left[\frac{\pi}{2} + \sin^{-1}(R\cos\phi)\right]\right\}$$

(3) EQUIPMENT

The transmitter used in this work was situated at Gib[raltar] (36° 8′ N, 5° 9′ W); it operated on a frequency of 37·3 Mc/s [with] a power of 40 kW. Two alternative horizontally-polarized [trans]mitting aerials were provided, one being a $\lambda/2$ dipole [and the] other an array having a plane-wave gain of 15 dB; both [were] 210 m above sea level. The polar diagram of the array [in the] horizontal plane has been measured by Crow *et al.*[3]; the dir[ection] of the middle of the beam is taken on the basis of this w[ork to] be 6° E of N, its width being about 25° between 3 dB p[oints.] During periods when ionospheric back-scatter[3] was likely [to be] received, pulses of length 10–20 ms were transmitted [with] frequency-shift keying; at other times when our observ[ations] were carried out the carrier was not modulated.

The receiver was situated near Slough (51° 31′ N, 0° 34′ W) [and] was in operation there from April, 1956, to January, 1957, [except] for a period in September, 1956, when the equipment was [moved] to Castlemartin (51° 38′ N, 5° 2′ W) in west Wales. The dis[tance] of Slough from Gibraltar is 1 740 km, and that of Castle[martin] about 20 km less. The centre of the beam of the transm[itting] array was directed approximately 5° W of the great-circle b[earing] to Slough and 6° E of the great-circle bearing to Castlema[rtin].

The signals were picked up at each receiving site on tw[o] dipoles arranged as described in Section 2 at a height of [...] The aerial outputs were taken to balanced receivers and [from] there to phase-measuring equipment of the sum-and-diff[erence] type.[1] To ensure gain and phase balance in the appar[atus, a] separate aerial was placed symmetrically relative to th[e] receiving aerials, 27 m from their mid-point and 3 m high. [This] aerial could be made to radiate 37·3 Mc/s, providing [equal] signals at each aerial for lining-up purposes.

To take observations, the aerials were first set a wave[length] apart, the equipment was lined up, and the gain set so [that a] signal at the median level would occupy roughly one-thi[rd of] full scale on the phase-display tube. The phase diff[erence] between the aerial outputs was then recorded photograph[ically] four times a second for a suitable time. This process [was] repeated for all other desired aerial spacings up to 10[λ, the] overall time taken from 20 to 60 min. The time of observ[ation] at each spacing was increased from ½ min at λ to 5 min at [...] compensate for the reduced accuracy of estimating the [para]meters of the widely-spread distributions expected at [large] spacings. The aerials were slung from ropes to enable [their] position to be altered quickly. To study the changes in pr[opaga]tion conditions occurring during a day, observations [were] generally made at a fixed aerial spacing of λ. Mean be[arings] were deduced from these fixed-spacing results by mea[ns of] eqns. (4) and (5).

As a check on the equipment and site, calibrations with [an] oscillator were carried out at each receiving point at a di[stance] of about 400 m from the aerials. With an aerial spacing [...] the average bearing error as deduced from the measured [...] was less than 1° for both sites.

To provide another test of the accuracy of the measur[ements] made with spaced aerials, some observations of the mean b[earing] of arrival of forward-scattered waves were also carried ou[t]

ssed-dipole aerial system. Of 22 pairs of mean bearings, h of which was taken within 10 min of the other, it was found t the differences between the bearings indicated by the ced-aerial and the crossed-dipole systems had an overall an of $-0\cdot5°$; this is quite satisfactory. The mean deviation m the mean in these differences was $2\cdot3°$.

REDUCTION AND ANALYSIS OF THE OBSERVATIONS

(4.1) Reduction

The results of the work were taken from the equipment in form of photographic records. After processing, these re placed in a film reader and the angle between the major s of each ellipse and a fixed direction was read off to give the antity $\psi/2$, where ψ is the phase difference required. If the ce amplitude exceeded full scale on the phase display, no ding of phase was taken from it, as a signal of such strength s assumed to be due mainly to one comparatively intense teor trail; such a condition is not compatible with the theory d here and the case was therefore excluded. Also, if the dian signal itself exceeded a particular value, the propagation s assumed to be due to reflection from a sporadic-E layer her than to ionospheric scatter and no observations were made. e quantities $\psi/2$ were punched directly on cards which were red until enough had been accumulated to be dispatched to National Physical Laboratory for computation of ν and R each group of values [eqns. (1) and (2)].

In dealing with fixed-spacing results the mean bearing was rived from the value of ν for each group by means of eqns. (4) d (5). A parameter is also required which will act as a asure of the spread of energy in the incident radiation. For s purpose each angular spectrum has been assumed to be ussian in form and its standard deviation has been chosen to fil this function; it can be calculated from the value of R by ans of the equation

$$R = \exp(-2\pi^2\sigma^2y^2) \quad . \quad . \quad . \quad . \quad . \quad (9)$$

ere σ is the standard deviation in question.

To produce an angular power spectrum from a series of ups of phase differences with different aerial spacings, the lowing procedure was adopted. The values of ν and R at the integral values of y from 0 to 10 were noted. If results one or more spacings were missing from the series, values re interpolated from the two nearest spacing results. Missing ults were of common occurrence, since the transmission edule often only allowed half-an-hour for the whole series d it was difficult to cover every aerial spacing in a whole hour, ecially at night. The integration involved in eqn. (3) was n carried out by the simple trapezoidal rule, except that the ction $R\cos\nu$ was considered to be parabolic in the region m $y = 0$ to 1, with zero slope at the origin. For values of greater than 10, R was taken to be zero.

In the above, R and ν for a spacing y are effectively the magude and phase of the Fourier component of the angular ctrum $Q(s)$ of period in s equal to $1/y$. As the spacing $= 10$ was the largest used, this means that details separated much less than $0\cdot1$ in s cannot be resolved in this way. In ctice, as the component for $y = 10$, which is of the form $(20\pi s)$, falls from a maximum to zero over a range of $0\cdot025$ s, it might be thought reasonable to give points of $Q(s)$ arated by this amount. However, as $y = 10$ was not always d, the results for $Q(s)$ have been calculated at a separation in f $0\cdot04$.

(4.2) Validity and Accuracy of the Methods Used

The method of deriving the angular spectrum can fail to give urate results in two main ways—first, if the assumptions made in the theory are invalid, and secondly, if the angular spectrum does not remain sufficiently constant over the period of observations.

To test the validity of the theory used, a number of distributions of phase difference about their mean were examined to see if they conformed to the distribution given by eqn. (8). Of the 12 cases tested, two gave distributions which were significantly different from the theoretical one at the 1% probability level. One of the other differences was significant at the 5% probability level. This certainly suggests that the assumptions made in the theory are not always valid. It may be noted here that this is contrary to the conclusions reached by Hagfors and Landmark,[5] who studied the dependence upon amplitude of certain functions of phase difference. Examples of the actual and theoretical phase distributions obtained at Slough are given in Fig. 1: (*b*) shows a significant difference between the two, but (*a*) does not.

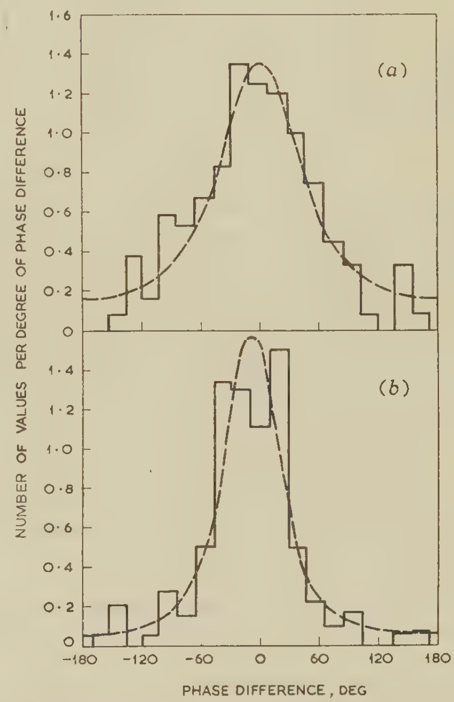

Fig. 1.—Phase-difference distributions.

Aerial spacing, one-wavelength.
(*a*) 2205, 30th November, 1956.
(*b*) 0612, 30th November, 1956.
——— Observed.
– – – – Calculated from eqn. (8).

It was considered advisable to pursue this matter further and so a comparison was made of the values of amplitude correlation coefficient, ρ_A, derived in two different ways. One was obtained from the signal amplitudes while the other was derived by means of eqn. (6) from the value of R which, in turn, was obtained from the phase-difference distribution by eqn. (2). The signal amplitudes at the two aerials are related to the lengths of the major and minor axes of the ellipse in the phase display. Suppose that these lengths are p_k and q_k respectively and that $k = 1, 2, 3, \ldots n$.

It is shown in Section 11.3 that the amplitude correlation coefficient ρ_A is given by

$$\rho_A = \frac{\Sigma p_k^2 - \frac{1}{n}(\Sigma p_k)^2 - \Sigma q_k^2}{\Sigma p_k^2 - \frac{1}{n}(\Sigma p_k)^2 + \Sigma q_k^2} \quad . \quad . \quad . \quad (10)$$

Some of the results obtained in this way are shown in Table 1, where $0{\cdot}91R^2$ is the value of amplitude correlation coefficient obtained from the phase results, and ρ_A is that from the amplitude measurements. According to eqn. (6) these should be the same for values of R not near to unity, provided, of course, that the theory is completely valid.

Table 1

COMPARISON OF PHASE AND AMPLITUDE RESULTS FOR THE AMPLITUDE CORRELATION COEFFICIENT

Date (1956)	U.T.	y	$0{\cdot}91R^2$ R from eqn. (2)	ρ_A from eqn. (10)
9th May	2235	1	0·63	0·70
9th May	2243	2	0·28	0·37
9th May	2251	3	0·04	0·42
16th August	0434	6	0·02	0·67
23rd August	1559	6	0·03	0·56
24th August	1524	6	0·00	0·69
24th August	1532	9	0·01	0·45
24th August	1856	6	0·05	0·33
1st December	0438	9	0·00	0·33

It will be seen that the agreement between the two is very poor, except for the narrow aerial spacings with which are associated the higher values of $0{\cdot}91R^2$. To explain this result, an effect must be sought which would tend to make peaks (or minima) of signal occur simultaneously at the two aerials in use, but which would still give a wide spread of phase differences. Now this is just what would happen if individual strong meteor signals were not eliminated satisfactorily from the records. For if, at any instant, a particular meteor signal increased so that it was larger than all other signals combined, then a maximum of signal due to it would frequently occur at the same time at both aerials, but there would be no noticeable tendency for this to occur near any particular value of phase difference for the large aerial spacings.

It may be noted here that ρ_A was brought nearer to $0{\cdot}91R^2$ by reducing the rejection level for strong signals, but they could not be made to agree without removing a large percentage of the initial data.

It must now be clear that the initial assumptions behind the theory cannot be regarded as entirely valid, since the high amplitude correlations observed directly at wide aerial spacings preclude this. The hypothesis that the discrepancies are caused by the presence of meteor reflections in the received signal is strengthened by an examination of the case of phase distribution (with narrow aerial spacings) which did not agree well with the original multi-signal theory. The distribution in Fig. 1(*b*) is one of these, and it shows a dip near 0° phase difference, which corresponds to the great-circle bearing of Gibraltar. It is a well-known result in the theory of reflections by meteor trails[6] that few reflections will be obtained from the near neighbourhood of the great-circle direction.

The question must now be considered of whether the failure of the original assumptions invalidates all the results obtained. In the first place, consider the results for mean bearing and energy spread derived from fixed-spacing measurements as explained in Section 4.1. Take an extreme case of the meteor effect by supposing that the entire signal is due to single plane waves building up and fading out rapidly in such a way that only one is present at any instant. Here the mean bearing of the incident energy, averaged over time, is still quite reasonably calculated by eqns. (4) and (5) since it will make little difference whether the averaging process of eqn. (1) is applied to the bearings or the phase differences; all results based on bearings can therefore still be taken to be correct. How for this situation the spread of the energy will be rather diffi It is simpler here to deal with the mean deviation from the of the phase-difference distribution, rather than R. Tl defined by

$$\beta = \frac{1}{n}\sum_{k=1}^{n}|\psi_k - \nu + 2m\pi|$$

so that $R = \cos\beta$. Then a Gaussian incident distributi energy in azimuth will have a standard deviation, σ_1, given

$$\sigma_1 = \frac{\beta}{2\sqrt{(2\pi)}y} \quad . \quad . \quad . \quad . \quad . \quad .$$

for the case of single incident plane waves. The compa formula for the multi-signal case can be obtained from eq and is

$$\sigma = \frac{\sqrt{\log \sec \beta}}{(\sqrt{2})\pi y} \quad . \quad . \quad . \quad .$$

Thus for an assumed Gaussian distribution of en eqns. (11) and (12) give the standard deviation on the ba the two theories. A plot of σ and σ_1 against β is show Fig. 2. The difference between σ and σ_1, though apprec

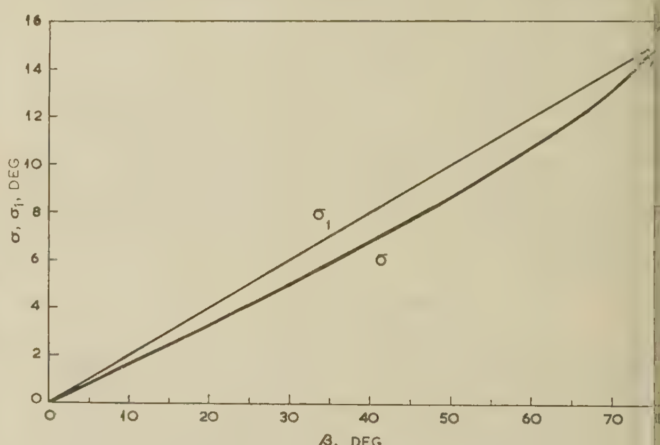

Fig. 2.—Variation of σ and σ_1 with β.
See eqns. (11) and (12).

is not great; in going over to the single-signal theory the bi change involved is an increase of 1·3° to the standard devi in the region near $\beta = 40°$. As the true situation probabl between the extremes of the multiple-signal and single-s theories, it is probable that an error of less than 1° in σ w caused by using the results of the multiple-signal theory ap to any fixed-spacing measurements.

The situation is more uncertain where the results are consi of the variable-spacing work, which should give the cor angular spectrum on the multi-signal theory. Most o phase-difference distributions examined are in agreement this theory, and it therefore seems justifiable to present the results obtained in this way. The satisfactory agree between theory and experiment found by Hagfors and Land also lends support to this course of action. However, clear from the considerations already discussed in this Se that the results cannot be considered to be as well establ as those obtained from the fixed-spacing measurements; are therefore given in separate Sections in the following.

To study the other source of error in this work—the cha in propagation conditions during one hour—tests were mad

amining the values of ν and R obtained at a fixed spacing (of .e or two wavelengths) at 5 min intervals. If the results are anged into terms of bearing, the distribution of mean bearings tained in this way have, for example, a standard deviation 0·9° centred on an overall mean of 3·2° W, and the values standard deviation which represent the spread of the energy azimuth have an s.d. of 0·4° and a mean of 4·4°. These sults appear reasonably steady and indicate that propagation nditions are usually sufficiently stable for this purpose. As additional precaution, those variable-spacing results which incided with a marked change of amplitude level during the riod in observations were rejected.

5) MAIN RESULTS AT SLOUGH AND CASTLEMARTIN

(5.1) Results with Fixed Aerial Spacing

The results which are discussed in this Section were obtained th the use of aerials at a fixed spacing, usually of λ. The ost striking discovery made by this means was the variation mean bearing with time of day, which was shown by the eliminary results at Slough reported in the previous paper.[1] ne complete set of values at Slough is shown in Fig. 3, which

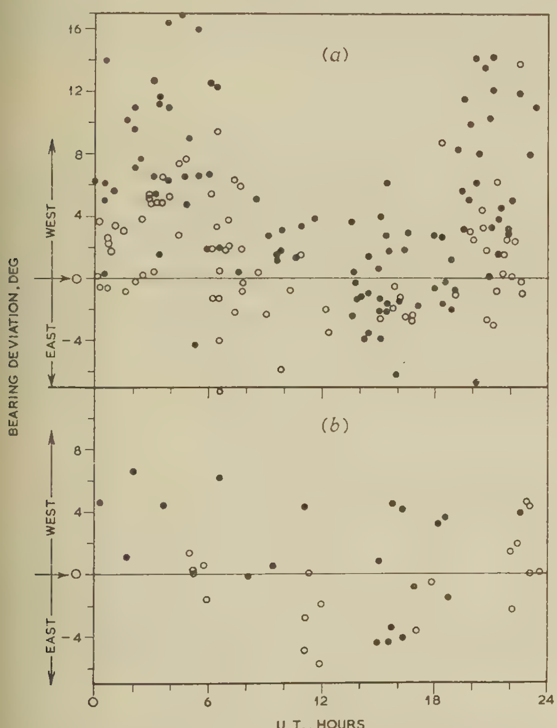

Fig. 3.—Diurnal variation of the mean bearing of signals received at Slough from Gibraltar by ionospheric forward scattering.

● Excluding December.
○ December only.
(a) Transmitter array.
b) Transmitter dipole.

cludes in Fig. 3(b) results when the transmitting aerial was a pole. In these Figures, all the points included refer to times parated by at least 20 min to avoid undue weight being given temporary abnormal conditions. As bearings taken in or ear December appeared to be somewhat different from the rest, ey have been shown with a separate symbol.

In Fig. 3(a) it will be seen that the mean bearing is considerably west of the great-circle path by night, whereas on the average is nearly on course during the day. In the December results, however, the westward deviation at night is not nearly so marked. If they are excluded, the mean value for the mean bearing in the period of day from 2100 to 0600 is 7·8° to the west, while for the period 1200–1800 this mean value is 0·4° to the east. The dipole results in Fig. 3(b) are similar, although the difference in the bearings is less marked; the corresponding figures are 4·2° to the west for the period 2100–0600 and 1·0° to the east for 1200–1800. It will be noted that there is a good deal of scatter about the mean value in the results.

To give some idea of the spread of the incident energy in azimuth for each group of observations, the standard deviation, σ, of the angular power spectrum has been calculated from eqn. (9) for each case and the results are plotted against time of day in Fig. 4. The results for the transmitter array and

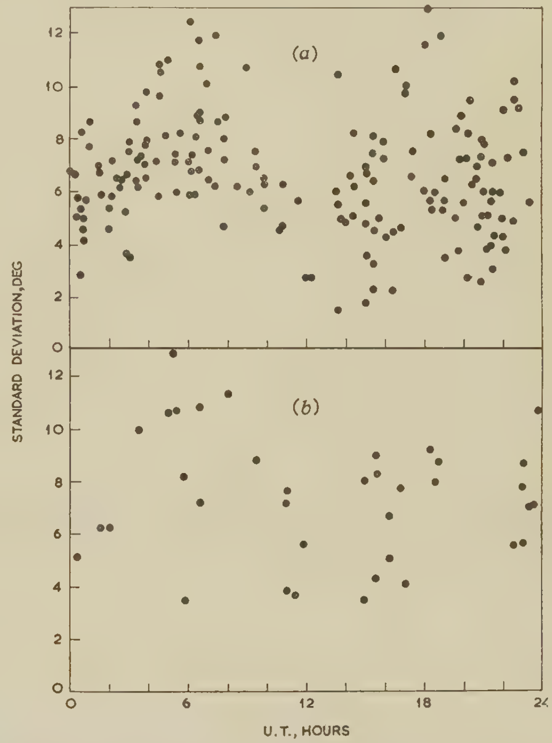

Fig. 4.—Diurnal variation of the standard deviation of the azimuth distributions of energy received at Slough from Gibraltar by ionospheric forward scattering.

(a) Transmitter array.
(b) Transmitter dipole.

dipole aerials are shown separately, but the December values are not specially indicated since they did not appear to be noticeably different from the others. There is again much scatter in the results, but if the mean value of σ is calculated for the transmitter array in use for each of the 3-hour periods, 0000–0300, 0300–0600, etc., maxima are found for the periods 0600–0900 and 1800–2100 of 8·2° and 6·6° respectively. Maxima for these periods are also shown in the transmitting dipole results, being of amplitude 9·8° and 8·6°; indeed, the dipole results are very similar to the array ones, though with slightly higher values of σ on the whole.

It will already be obvious that there is some seasonal variation in the results from the difference shown by the December values. Plots of the results against time of year have been made for bearing in 3-hour groups during the day, but these show little more than the absence of large westerly bearings on December nights, so they are not reproduced here. There is some indication, however, that this effect is part of a cyclical change during

the year, with maximum westerly bearings at night in the period August–October, and minimum in December; unfortunately, no experimental data are available from early January until the beginning of April.

The results obtained in September, 1956, with a receiver in Castlemartin on the westward side of the main lobe of the Gibraltar array must now be considered. A plot of the mean bearing against time of day is shown in Fig. 5. The results for the transmitter array show a considerable eastward trend

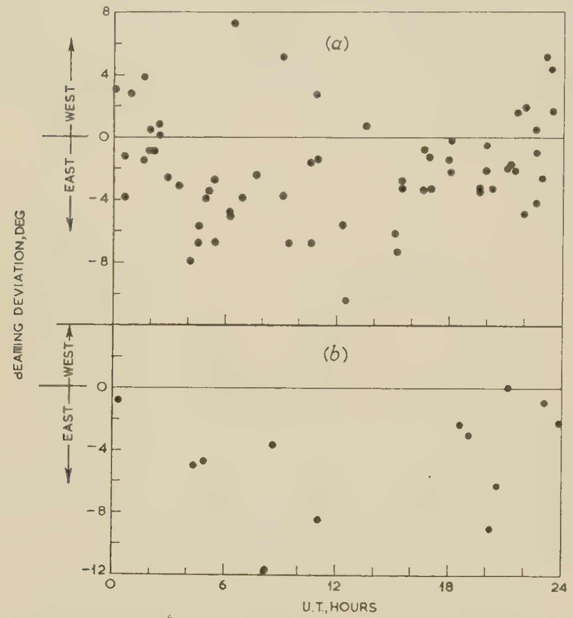

Fig. 5.—Diurnal variation of the mean bearing of signals received at Castlemartin from Gibraltar by ionospheric forward scattering in September, 1956.

(a) Transmitter array.
(b) Transmitter dipole.

during the day, but the mean bearing is nearly on the great-circle direction during the night, at least from 2100 to 0... This result is nearly the reverse of what is found at Slough except that the bearings during the day at Castlemartin are so far off course as those at Slough during the night. ... bearings with the transmitter dipole in use are few in number but nevertheless are surprisingly similar to the array results; ... point will be discussed in Section 7.

It is very instructive to compare bearings obtained on forward-scatter signals with those on individual comparatively strong meteor reflections. Practical results for the Gibraltar–Slough path had previously been obtained by Meadows,[7] and an effort was therefore made at Castlemartin to obtain some records of meteor bearings during the experiments there. This was done by reading off the phase differences given by these meteor signals on the existing equipment during the intervals of the other work. These phase differences were converted to bearings, and Table 2 gives the bearing distributions obtained from them in the various 3-hour periods during the day. Three-hour periods with no results are omitted, and mean bearings for the distributions are given only where the numbers are sufficiently large to justify them. The transmitter-array results show the same trend as the forward-scatter mean bearings with even more decidedly eastward values during the day. The transmitter-dipole results contain the most westerly bearings obtained, ... the mean for the period 1800–2100 is still east of the great-circle direction of the transmitter.

(5.2) Results with Variable Aerial Spacing

The method of calculating the angular power spectrum from the variable-spacing results has been explained in Section ... An example of the values of R and ν obtained in a particular series of variable measurements is shown in Table 3. The values were converted to give the angular power spectrum shown in the solid curve of Fig. 6, which is the function $Q(s)$ plotted as a function of s, and is therefore automatically normalized [see eqn. (17)]. This particular curve is an interesting one as ...

Table 2

METEOR BEARINGS AT CASTLEMARTIN RELATIVE TO GREAT-CIRCLE DIRECTION

Bearing	Numbers of bearings in a 3° range								
	With transmitter array aerial							With dipole	
	0000–0300	0300–0600	0600–0900	1200–1500	1500–1800	1800–2100	2100–2400	1500–1800	1800–2100
27° E	1	—	—	—	—	—	—	—	—
24° E	—	1	2	—	—	—	—	—	1
21° E	1	—	1	—	—	—	—	—	1
18° E	3	1	2	—	—	—	—	—	—
15° E	0	7	6	4	—	—	—	—	1
12° E	6	11	2	3	—	—	1	1	2
9° E	3	9	4	1	2	—	1	1	2
6° E	4	5	2	6	0	—	3	2	2
3° E	1	2	1	5	8	3	1	1	1
0° E	9	4	3	2	3	1	11	—	7
3° W	14	10	4	—	—	1	15	—	2
6° W	12	7	—	—	—	—	4	—	1
9° W	5	3	—	—	—	—	1	—	2
12° W	—	—	—	—	—	—	—	—	1
15° W	—	—	—	—	—	—	—	—	1
18° W	—	—	—	—	—	—	—	—	—
21° W	—	—	—	—	—	—	—	1	—
Mean bearing	1·2° E	4·9° E	8·8° E	7·4° E	3·2° E	—	1·0° W	—	2·5° E

Table gives numbers of bearings in a 3° range centred on the value given in the first column.

Table 3

VALUES OF R AND ν FOR 0638–0720, 17TH AUGUST, 1956, AT SLOUGH

y	R	ν deg
1	0·74	74
2	0·23	−16
3	0·08	13
4	0·06	−68
5	0·32	−10
6	0·03	154

ppreciable part of the incident energy comes from near the reat-circle direction, but some also comes from much more esterly bearings. Another example of an angular spectrum is iven in Fig. 7—this time it is of a more ordinary kind, showing ne strong peak slightly displaced to the east. The negative

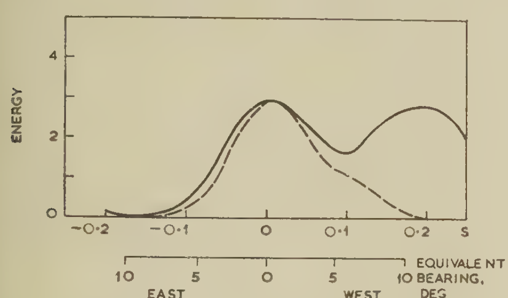

Fig. 6.—Angular power spectrum at Slough for 0638–0720, 17th August, 1956.

Transmitter aerial, dipole.
——— Energy arriving at site.
– – – – Energy picked up by rhombic.

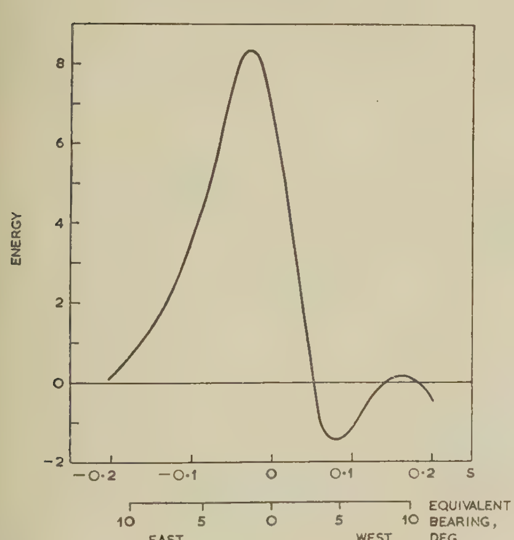

Fig. 7.—Angular power spectrum at Slough for 1106–1142, 8th January, 1957.

Transmitter aerial, dipole.

alues for the incident energy near $s = +0\cdot1$ are, of course, hysically meaningless. They do occur sometimes in the results, resumably owing either to errors in the initial data or to the umerical integration procedure not being sufficiently accurate.

To enable some conclusions to be drawn from the entire ollection of angular spectra available at Slough, the average alue of Q(s) was taken in each case for the ranges of s of $-0\cdot14$ to $-0\cdot10$, $-0\cdot02$ to $+0\cdot02$ and $0\cdot10$ to $0\cdot14$, corresponding to bearing ranges of 8·0° E–5·7° E, 1·1° E–1·1° W and 5·7° W–8·0° W. It was considered that the value of Q(s) around $s = 0$ would provide information about the energy propagated by a turbulent-scattering process, and the other values would do the same for energy reflected from meteor trails. Theoretically the area under Q(s) is always constant and equal to unity and so the quantities considered here are really percentages of the total energy and not absolute values. The mean of the average values over all spectra was taken for each range; day and night were treated separately, and so were the December results. Thirty curves for Q(s) were dealt with in this way; five were rejected for this analysis because the area under the curve appreciably exceeded unity. The results are given in Table 4.

Table 4

MEAN AMPLITUDE OF ANGULAR SPECTRA AT SLOUGH IN CERTAIN RANGES OF s

Time	Mean amplitude over ranges of s of		
	−0·14 to −0·10	−0·02 to 0·02	0·10 to 0·14
0600–1800, excluding Dec.	1·7	3·7	1·8
1800–0600, excluding Dec.	1·2	2·7	3·9
0600–1800, 29th Nov.–8th Jan.	2·0	4·6	1·5
1800–0600, 29th Nov.–8th Jan.	1·2	2·9	2·3

The transmitter array and dipole results have not been treated separately. Negative values of s correspond to easterly bearings.

The amplitudes in the westerly range 0·10–0·14 (excluding December) are significantly greater by night than by day, as would be expected if their were a considerable meteoric contribution to the signal.[7] The difference, though it still exists, is not statistically significant for December. Note, too, that the amplitude during the main part of the year is larger at night in the west (3·9) than in the on-course range (2·7). Another feature of Table 4 is the significant difference between day and night values of amplitude in December for the range around $s = 0$; this shows that the proportion of energy near the great-circle direction is greater by day than by night. On the whole these figures, which refer to Slough only, agree well with the fixed-spacing results. In particular, they confirm that a greater percentage of the energy comes from the west at night and that this tendency is less marked in December.

A number of variable-spacing observations were carried out at Castlemartin, one of them being that shown in Fig. 8. The

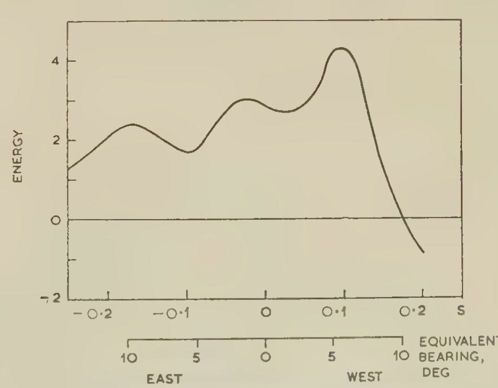

Fig. 8.—Angular power spectrum at Castlemartin for 0306–0350, 25th September, 1956.

Transmitter aerial, array.

calculations corresponding to those in Table 4 were also performed for the Welsh data and are given in Table 5.

It must be pointed out here that these daytime values for Castlemartin do not agree particularly well with what would be

Table 5

MEAN AMPLITUDE OF ANGULAR SPECTRA AT CASTLEMARTIN IN CERTAIN BEARING RANGES

Time	Mean amplitude over ranges of *s* of		
	− 0·14 to − 0·10	− 0·02 to 0·02	0·10 to 0·14
0600–1800	0·9	6·0	1·4
1800–0600	2·1	4·0	2·4

expected from the fixed-spacing results, which showed a marked eastward trend during the day. However, the sample here is rather small, as only four sets of daytime results are available, three of which were taken on one day.

(6) AERIAL-GAIN COMPARISONS AT SLOUGH

(6.1) Measurements with Fixed Aerial Spacing

A complete knowledge of the angular power spectrum at a point on the ground is necessary to enable one to calculate precisely the gain realized by any particular receiving aerial situated at that position. However, with the assumption that the shape of this power spectrum is Gaussian, the gain can be calculated from the fixed-spacing results alone; the accuracy of this deduction will, of course, depend to some extent on the goodness of this assumption, but it should not be too unsatisfactory.

A rhombic receiving aerial was in use at Slough during the period in question for the measurement of the signal strength of the Gibraltar transmissions. The input to the receiver could also be switched to a dipole aerial. For a period in 1956 the rhombic and dipole inputs were changed automatically every 10 min, thus providing a means of finding the gain of the rhombic over the dipole.* For all occasions when this switched-aerial reception was in use at the same time as the fixed-aerial-spacing measurements, already described, were taken, gains were computed by both methods; the results are shown in the scatter

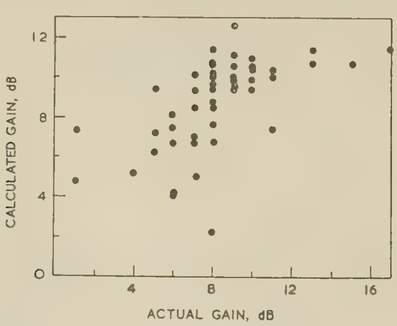

Fig. 9.—Comparison of actual and theoretical effective gains of the rhombic aerial at Slough and a dipole.

diagram in Fig. 9. To check the gains and polar diagrams of the Slough aerials, an extensive series of measurements was carried out in an aircraft from the Royal Aircraft Establishment, Farnborough. The shape of the main lobe of the rhombic

* This work was under the direction of Mr. G. W. Luscombe.

polar diagram turned out to be very close to that predicte[d by] theory, but its actual plane-wave gain over the dipole was ra[ther] lower than expected, namely 13 instead of 15 dB. The h[alf-] power points were at about ±6° from the maximum.

The relation between the actual and calculated gains show[n in] Fig. 9 does not show a perfect correspondence between the [two,] but the general tendency for each to increase together is c[lear-ly] evident. It will be noted that the actual gains are, on aver[age,] rather lower than the predicted ones by about 1 dB. [No] explanation for this effect has been found, although the [dis-]crepancy would, of course, be eliminated if the plane-wave [gain] of the rhombic were taken to be 12 instead of 13 dB; the di[rect] measurements from an aircraft might just be stretched to a[llow] this interpretation, but 13 dB seems the fairer value to [take] from them.

(6.2) Measurements with Variable Aerial Spacing

Theoretically, the fixed-spacing results cannot lead to su[ch] satisfactory calculation of the gain realized by any recei[ving] aerial as can the variable-spacing results, which, of course, [give] the angular power spectrum, but this factor is offset in [the] present work by the decreased reliability of the variable-spa[cing] experiments. However, if the angular power spectrum is re[ally] known, the power arriving from any direction has only to [be] multiplied by the receiving-aerial polar diagram to give [the] power reaching the receiver from that direction. The contri[bu-]tions from all directions can be added up and compared with [the] total power in the signal picked up by a $\lambda/2$ dipole. Fig[. …,] which shows a particular angular spectrum, also contain[s a] curve giving the reduction of power in each direction caused [by] the polar diagram of the rhombic aerial at Slough. Meas[ure-]ment of the area under this curve leads to the result that [the] gain of the rhombic over the dipole at this time would be 8·8 [dB].

The gain of the rhombic aerial has been calculated in this [way] for all of the variable-spacing results during which ac[tual] measurements of the gain were carried out. A compariso[n of] the values obtained in each way is given in Table 6. Un[for-]

Table 6

COMPARISON OF THE OBSERVED GAIN OF THE SLOUGH RHOM[BIC] AND THAT CALCULATED FROM THE ANGULAR PO[WER] SPECTRUM

Date (1956)	U.T.	Observed gain	Calculated gain
		dB	dB
17th Aug.	0700	8	8·8
21st Aug.	0830	10	8·3
22nd Aug.	1120	9	10·2
23rd Aug.	1550	11	10·1
24th Aug.	1515	7	10·2
24th Aug.	1850	10	8·4
17th Dec.	1920	10	11·4

tunately, not many results are available as the switching f[rom] rhombic to dipole was not carried out until August, 1956, [and] also because ionospheric back-scatter prevented the use of m[any] of the daytime measurements.

The two values are generally of the same order of magni[tude,] but the agreement in detail is not good. As the values sh[ould] certainly be the same, some errors must have arisen in [the] estimation of one or other of the values. It has already [been] pointed out that the variable-spacing results are not a[bove] suspicion, but it must also be said here that on many of [the] amplitude records (from which the observed gain was deri[ved)] the accuracy of reading the gain is not even as good as ±1

(7) DISCUSSION

Before the implications of the work considered in the previous ·ctions are examined, it is interesting to see what can be deduced om aerial-gain measurements by themselves when they are udied from the standpoint of the azimuth power distribution hich gives rise to them.

Comparative aerial-gain figures showing large gain degrada- on are often quoted, but it may not always be realized how nall a fraction of the total incident energy must lie near the eat-circle path for this to be possible. Consider, for instance, e rhombic aerial at Slough which has already been mentioned, ith half-power points at about $\pm 6°$. The effective gain of is aerial is less than a value between 8 and 9 dB for 50% of the me (Luscombe, unpublished results for August and September, 955). Now an angular power spectrum of Gaussian form, mmetrical about the great-circle path, would need to have alf-power points as much as 18° from its mean to give an fective gain as low as 8 dB, and, of course, still lower gains are und at times.

made do not appear likely to affect the answer seriously, and it is therefore clear that turbulent scattering from many sources simultaneously cannot alone account for the results on this particular path.

The results on energy distributions in azimuth which have been given in this paper show what is happening to give these low aerial gains. Usually it is because a considerable amount of energy has been displaced to one side of the path beyond the region of strong reception for the Slough rhombic aerial, as in Fig. 6. There is a good deal of variability in the results, however, and a mean bearing of zero is sometimes obtained. Now the nighttime results for Slough reception in particular suggest the likelihood of reflections from meteor trials being responsible for much of the energy reaching the receiver. As an additional check, a comparison has been made in Table 7 between mean bearings obtained in forward scatter and those found in reflections from discrete meteor trails; the latter values can be obtained from the data given by Meadows.[7]

In Table 7 mean bearings are in degrees and only results with

Table 7

Mean Bearings of Forward Scatter (from Present Results) and of Meteor Reflections (from Meadows) as Received at Slough

Type of signal	Bearings							
	0000–0300	0300–0600	0600–0900	0900–1200	1200–1500	1500–1800	1800–2100	2100–2400
	deg	deg	deg	deg	deg	deg	deg	deg
Forward scatter, excluding December	7·7	8·4	5·9	2·2	−0·7	0·4	4·7	6·6
Forward scatter, December only	2·2	5·4	2·0	—	—	−2·3	2·1	1·7
Meteor reflections	7·9	6·8	0·0	−0·4	−0·0	−1·5	4·9	7·3

Bearings are in degrees relative to the great-circle path, negative values representing easterly bearings.

This type of result leads one to wonder whether any type of urbulent-scattering mechanism can give a satisfactory explana- ion. Consider, for example, times when the transmitter array in use, and suppose as an extreme case that the amount of cattered power is independent of scattering angle. Now the mount of power returned from any azimuth can be calculated the same way as the number of meteor reflections from any zimuth was calculated in a previous paper,[4] and with the same pproximations. It is found in this case that the effective aerial ain would be 8·4 dB. Thus, even with this wide-angle type of cattering, the gain only falls to a value which is exceeded half he time, and the hypothesis fails to account for the still lower ains which are found for the rest of the time.

If a more plausible scattering law is taken, the turbulent- cattering theory is still farther from being able to explain the esults. Suppose that the scattering angle is X. Then the theory iven by Villars and Weisskopf leads to the result that the eceived power is proportional to $\sin^5 (X/2)$. The gain cal- ulations can be repeated with this scattering law incorporated nd this leads to the conclusion that the effective aerial gain vould be 11·2 dB—not nearly a small enough value. It is true hat certain approximations have been made in the calculations vhich may not be strictly true. For instance, the height of 00 km which was chosen for the scattering region may be too reat. However, the effect of reducing it would be to narrow he azimuth distribution of energy reaching the receiver and hus to increase the effective gain. Or the scattering might have een taken to occur over a range of heights, say from 80 to 00 km, but this would add more energy near the great-circle ath and give increased gain once more. The other assumptions

the transmitter array in use are included; they are based on all values obtained, grouped in 20 min periods, and so differ slightly from those in Fig. 3.

If the received signal were due to turbulent scattering alone, with a 5th-power scattering law, the mean bearing of the forward scatter would be 0·8° W when the transmitter array was in use. This figure differs from zero because of the westward displacement of the transmitter main lobe from the line Gibraltar to Slough. Hence, if an appreciable contribution of signal from turbulent scattering existed at any time, this would be shown by the mean bearing of the forward scatter (as given in Table 7) not being the same as the meteor mean (also given in Table 7) but lying nearer to 0·8° W. No very marked case of this nature is found in the results excluding December, the largest such effect being in the period from 1500–1800, where the forward-scatter value is 0·4° W and the meteor mean is 1·5° E; but this difference could not be called significant. There is a very noticeable discrepancy between the values for the period 0600–0900, but it is in the opposite direction. The reason for this is not definitely known, but it may be that the entire period 0600–0900 is not adequately covered in the samples taken. This idea is supported by the fact that the mean meteor bearing derived theoretically from a uniform heliocentric distribution of meteor radiants is 3·5° for this time, in contrast to the observed value of 0·0°; this distribution is known from other work[4] to be quite a good representation of the meteor results as a whole.

The general agreement between the mean bearings of meteor reflections and forward scatter at Slough indicates that during the night from April to November the forward-scatter signal is

almost entirely due to reflections from meteor trails. However, as the meteor bearings themselves have values nearly along the great-circle to Gibraltar during the day, daytime observations at Slough cannot be expected to distinguish between the two possible mechanisms for producing forward scatter. The asymmetrical property of the mean meteor bearings at Slough, namely that they lie on the great circle by day and are 7° W of it by night, is largely due to the 5° westward deviation of the main lobe of the transmitter polar diagram. Therefore, if observations are taken at a point which is at a comparable angle in azimuth on the other side of the lobe, one would expect that the average meteor bearing would be roughly along the great circle to Gibraltar during the night and 7° eastwards of it during the day. This expectation is indeed borne out by the results of the meteor-bearing experiments at Castlemartin which are quoted in Table 2. But if we turn to the forward-scatter results given in Fig. 5 for the same receiving point, it will be seen that during the day only one or two mean bearings are as far deviated as 7° E, the mean being nearer 3° E. The mean bearing given by turbulent scattering alone would be 0·8° E, and so, as the actual mean lies somewhat nearer to this figure than to 7° E, it must be concluded that turbulent scattering is of considerable importance during the daytime, contributing on the average rather more than half the power in the total signal.

The situation is not so clear at the time of year around December, since there are no observations at Castlemartin for this time. However, the results do suggest that at night at Slough there is an appreciable amount of signal coming in from near the great-circle direction, and thus that there is a substantial contribution to the energy from turbulent-scattering processes. This may explain, at least in part, the fact that the strongest mean signals of the year occur in December (see, for instance, Isted,[9] who received the signal near Chelmsford on an array with little directivity in azimuth).

The results obtained at Slough when a dipole aerial was in use for transmitting agree reasonably well with the above conclusions. The eastward deviation of mean bearing during the day is not nearly as large as would be expected from meteor reflections alone, although it is somewhat greater than with the transmitter array. Also, the westward deviation at night is not so marked. Both these results agree with what the broader polar diagram of a dipole would give.

A few results with the dipole transmitting aerial were obtained at Castlemartin. Now, with its wide beam, it ought to give much the same sort of azimuth power distribution at Castlemartin as at Slough. This is certainly not what is found, for the mean bearings are eastward by day and nearly on course at night; indeed, there is not even one example of a westerly mean bearing at night. The only plausible explanation seems to be that the radiated power from the dipole at Gibraltar falls off considerably to the west of bearing 0°, the bearing of Castlemartin. Fortunately there is a little corroborative evidence for this in some meteor bearings which were taken on the dipole transmissions at Castlemartin for the period 1800–2100. These show a few westerly meteor trails, but the mean bearing of the whole group is 3° to the east of the great-circle direction, instead of being several degrees to the west.

When these results are applied to the consideration of other ionospheric-scatter paths, it should be borne in mind that the meteor bearing distribution will be affected by the orientation of the path; however, this distribution is calculable, as was shown in a previous paper.[4] Also, it may well be the case that turbulent-scattering processes become of more importance as the path length decreases, for the extent of the scattering volume common to transmitter and receiver will then become greater in the lower regions of the ionosphere.[10] Meteors do not penetrate in such large numbers below 90 km in height, and effect of turbulence is expected to be greatest in the lower p of the E-region.

(8) CONCLUSIONS

On the paths studied here, of 1 740 km from Gibraltar, it been shown that at all times of day the distributions of ph and amplitude obtained with a pair of spaced aerials are s that the signal cannot be due to scattering from a large num of irregularities. It appears that a strong wave from direction is frequently predominant; this, however, is w would be expected from meteor reflections. The results obtai by calculating the angular power spectrum from measureme of phase differences at a number of aerial spacings are theref somewhat unreliable, since they are based on the assumption a large number of incident signals of random phase; but m bearings obtained from them are not affected by this. It also been shown that the results from observations with a si narrow spacing are still valid. These show a marked diur variation of mean bearing at Slough, about 7° W of the gr circle direction at night but near 0° W of it by day. At Cas martin, on the other side of the transmitter beam, the m bearing was along the great-circle direction by night and by day. By comparing these values with the results of mete reflection experiments, it was shown that this indicates that signal is almost entirely meteoric by night, but that during day turbulent scattering contributes somewhat more than the incident energy. Around December, turbulent scatter appeared to be appreciable even during the night.

(9) ACKNOWLEDGMENTS

The transmitter was operated by Marconi's Wireless T graph Company, Ltd., under contract to the Admiralty Ministry of Supply; the co-operation of these organization gratefully acknowledged.

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(11) APPENDICES

(11.1) Derivation of Eqns. (1), (2) and (3)

Bramley[2] considered the theory appropriate to the case of a continuous distribution in azimuth of incident waves. Suppose that $P(\alpha)d\alpha$ is the power arriving in the sector $\alpha \pm \frac{1}{2}d\alpha$. Then he defined two functions μ_{13}, μ_{14} such that

$$\mu_{13} = \int_0^{2\pi} P(\alpha)\cos(2\pi y \sin\alpha)d\alpha \quad . \quad . \quad . \quad . \quad (13)$$

$$\mu_{14} = -\int_0^{2\pi} P(\alpha)\sin(2\pi y \sin\alpha)d\alpha \quad . \quad . \quad . \quad (14)$$

Note that the α used here differs by $\pi/2$ from that used by Bramley. Then he defines

$$R^2 = (\mu_{13}^2 + \mu_{14}^2)/P_0^2 \quad . \quad . \quad . \quad . \quad . \quad (15)$$

where P_0 is the total incident power. Also,

$$\tan\nu = -\mu_{14}/\mu_{13} \quad . \quad . \quad . \quad . \quad . \quad (16)$$

From eqns. (13)–(16) it can be deduced that

$$R\exp(j\nu) = \frac{1}{P_0}\int_0^{2\pi} P(\alpha)\exp(j2\pi ys)d\alpha$$

putting $s = \sin\alpha$. Hence

$$R\exp(j\nu) = \frac{1}{P_0}\int_{-1}^{1}\frac{P(\sin^{-1}s)}{\sqrt{(1-s^2)}}\exp(j2\pi ys)ds$$

$$= \int_{-1}^{1} Q(s)\exp(j2\pi ys)ds$$

where
$$Q(s) = \frac{P(\sin^{-1}s)}{P_0\sqrt{(1-s^2)}} \quad . \quad . \quad . \quad . \quad . \quad (17)$$

Defining $Q(s) = 0$ for values of $|s| > 1$,

$$R\exp(j\nu) = \int_{-\infty}^{\infty} Q(s)\exp(j2\pi ys)ds \quad . \quad . \quad . \quad (18)$$

Both sides of this equation are functions of y and not of s, so that the expression is of a standard Fourier transform and can be inverted, giving

$$Q(s) = \int_{-\infty}^{\infty} R\exp(j\nu)\exp(-j2\pi ys)dy$$

$$= \int_{-\infty}^{\infty} R\cos(\nu - 2\pi ys)dy - j\int_{-\infty}^{\infty} R\sin(\nu - 2\pi ys)dy$$

Now $\int_{-\infty}^{\infty} R\sin(\nu - 2\pi ys)dy$

$$= \int_{-\infty}^{\infty}(R\sin\nu\cos 2\pi ys - R\cos\nu\sin 2\pi ys)dy$$

R is always positive and an even function of y, but ν is an odd function of y. Hence, both terms of the integral are odd functions of y, and its value must be zero. So

$$Q(s) = \int_{-\infty}^{\infty} R\cos(\nu - 2\pi ys)dy$$

$$= \int_{-\infty}^{\infty}(R\cos\nu\cos 2\pi ys + R\sin\nu\sin 2\pi ys)dy$$

$$= 2\int_0^{\infty}(R\cos\nu\cos 2\pi ys + R\sin\nu\sin 2\pi ys)dy$$

as both terms are even functions of y. This is eqn. (3).

It remains to indicate how eqns. (1) and (2) are derived. Eqn. (2) is readily obtained from Bramley's eqn. (36), but eqn. (1) is not found quite so obviously. The theoretical distribution of phase differences is symmetrical about a certain value, and we have to make ν the best estimate of this value from the information given in each sample. Some sort of mean is clearly required, but taking the arithmetic mean leads to difficulties associated with the fact that the phase 359° actually differs from the phase 0° by only 1°. The definition of ν selected here can be thought of as giving the direction of the centroid of the phase differences if these are arranged on the circumference of a circle. It is certainly an unbiased estimate, and also reduces to the arithmetic mean for a very narrow distribution not near 0°.

(11.2) The Value of ν if $Q(s)$ is Symmetrical

If $Q(s)$ is symmetrical about a certain value, s_0, of s, then

$$Q(s_0 - p) = Q(s_0 + p)$$

We have, from eqn. (18),

$$R\exp j\nu = \int_{-\infty}^{\infty} Q(s)\exp(j2\pi ys)ds$$

$$= \int_{s_0-s_1}^{s_0+s_1} Q(s)\exp(j2\pi ys)ds$$

assuming that $Q(s)$ is zero for values of s removed by more than s_1 from s_0. Changing the variable of integration,

$$R \exp jv = \int_{-s_1}^{s_1} Q(s_0 - p) \exp [j2\pi y(s_0 - p)]dp$$

$$= 2 \exp (j2\pi y s_0) \int_0^{s_1} Q(s_0 - p) \cos (2\pi y p)dp,$$

using the symmetry property. The integral now contains only real quantities, and hence

$$v = 2\pi y s_0$$

as quoted in eqn. (4).

(11.3) Deduction of Amplitude Correlation Coefficient from Measurements on the Phase Display

Suppose that a, b are the amplitudes of the signals arriving at the two aerials and define p, q as $p = a + b$ and $q = a - b$.

Let σ_a, σ_b, σ_p, σ_q be the standard deviations of the quantities denoted in the suffices. As conditions will be the same on average at each aerial we have

$$\sigma_a = \sigma_b \quad . \quad . \quad . \quad . \quad . \quad . \quad (19)$$

and

$$\frac{1}{n} \sum_{k=1}^{n} q_k = 0 \quad . \quad . \quad . \quad . \quad . \quad . \quad (20)$$

Eqn. (20) implies that the mean value of q_k may be taken to be zero; it must be assumed that n is large.

If r is the correlation coefficient between a_k and b_k, then by using the expressions for the variance of a sum and of a difference, we have

$$\sigma_p^2 = \sigma_a^2 + \sigma_b^2 + 2r\sigma_a\sigma_b$$
$$= 2\sigma_a^2(1 + r)$$

Similarly,

$$\sigma_q^2 = 2\sigma_a^2(1 - r)$$

Hence

$$\frac{\sigma_p^2}{\sigma_q^2} = \frac{1 + r}{1 - r}$$

and

$$r = \frac{\sigma_p^2 - \sigma_q^2}{\sigma_p^2 + \sigma_q^2}$$

We have now merely to express σ_p and σ_q in terms of values of p and q. This gives

$$r = \frac{\frac{1}{n}\Sigma p_k^2 - \left(\frac{1}{n}\Sigma p_k\right)^2 - \frac{1}{n}\Sigma q_k^2 + \left(\frac{1}{n}\Sigma q_k\right)^2}{\frac{1}{n}\Sigma p_k^2 - \left(\frac{1}{n}\Sigma p_k\right)^2 + \frac{1}{n}\Sigma q_k^2 - \left(\frac{1}{n}\Sigma q_k\right)^2}$$

$$= \frac{\Sigma p_k^2 - \frac{1}{n}(\Sigma p_k)^2 - \Sigma q_k^2}{\Sigma p_k^2 - \frac{1}{n}(\Sigma p_k)^2 + \Sigma q_k^2}$$

using eqn. (20).

This result has the same form as eqn. (10). Note that unaffected by changing the sign of q_k, and hence it still gives amplitude correlation coefficient if we define p, q by the relatio

$$\left.\begin{aligned} p &= a + b \\ q &= |a - b| \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad .$$

With this definition p, q can be taken to be the lengths of axes of the ellipse on the phase display, since Ross, Bram and Ashwell[11] gave for these lengths the relation

$$\frac{q}{p} = \frac{|1 - (b/a)|}{1 + (b/a)}$$

$$= \frac{|a - b|}{a + b}$$

If we ignore the constant of proportionality, which is of account in correlation work, this gives eqn. (21). Hence eqn. gives the relation expressing the correlation coefficient in te of the lengths of the axes.

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IRECTIONAL OBSERVATIONS ON DELAYED SIGNALS ON AN IONOSPHERIC FORWARD-SCATTER CIRCUIT

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SUMMARY

A study has been made of delayed signals reaching Slough from a .nsmitter at Gibraltar on 37 Mc/s in the winter of 1956–57. The ·an bearing of the first signal to arrive after that due to ionospheric ·ward-scatter was found to vary between 250° and 290° during the y; the spread of energy in bearing is also considered. It is shown ıt the radiation from the transmitter aerial must be stronger the west than to the south. Results for the second delayed ;nal are also given. Round-the-world echoes were occasionally served, but these did not arrive from the direction of the great circle ntaining Gibraltar and Slough.

(1) INTRODUCTION

Directional observations were carried out at Slough during the ·riod April, 1956, to January, 1957, on the forward-scatter ;nal from a transmitter at Gibraltar with horizontally-olarized aerials on the frequency of 37·3 Mc/s. This work, hich has been described in a previous paper,[1] was carried ıt by means of measurements of the phase difference between e signals arriving at a pair of aerials spaced at right angles the great-circle direction of Gibraltar. The work of Crow *al.*[2] had previously shown that interference from back-scatter gnals was likely to be experienced frequently; at times when is was expected the transmitter sent pulses of 10 ms length. his enabled the forward-scatter and back-scatter signals to studied separately, and the present paper describes directional easurements on the latter.

The measuring technique employing spaced aerials was not, owever, particularly suitable for obtaining bearings on the ıck-scatter. The polar diagram of the spaced dipoles fell off o much at angles far removed from the great-circle direction, ıd there were, in general, four ambiguities in bearing for a ven phase difference, even if the smallest possible aerial spacing f one wavelength) was used. It was therefore decided to use crossed-dipole system to measure azimuth; the aerials were orizontal at a height of 23 m and were each connected to one annel of a twin-channel receiver with a cathode-ray-tube splay. Such a system has a 180° bearing ambiguity and is, course, much more subject to polarization error than is the aced-aerial system, but it was considered adequate for the curacy required, which was of the order of ±5° for a mean earing.

Each received signal on the display had the form of an ellipse hich changed continually in size, shape and the direction of major axis. The last quantity is what would be considered e bearing of an ordinary signal, and will be referred to here the 'indicated bearing'. Now, a given distribution $P(\alpha)$ of cident energy (expressed as a function of α, the bearing in dians relative to some particular direction) will give rise to a rticular distribution $g(\theta)$ of indicated bearing θ. The form of θ) is not known for this particular problem, but if the dipole olar diagrams are taken to be sinusoidal in shape, and if the incident energy is restricted to a sector of about 50° in azimuth, it can be shown that

$$g(\theta) = \frac{\nu^2}{2(\theta^2 + \nu^2)^{3/2}} \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where

$$\nu^2 = \frac{\int_{-\pi}^{\pi} P(\alpha) \sin^2 \alpha d\alpha}{\int_{-\pi}^{\pi} P(\alpha) \cos^2 \alpha d\alpha} \quad . \quad . \quad . \quad . \quad . \quad (2)$$

In the above, α must be measured from the mean bearing of the incident radiation; ν can be seen from eqn. (2) to approximate to the standard deviation of the distribution $P(\alpha)$, as $P(\alpha)$ is assumed to be appreciable only when α is small. It is interesting to note that a relation of the same form as eqn. (1) was obtained by Whale and Delves[3] for indicated bearings taken on another type of narrow-aperture direction-finder; an account of the methods used in deriving distributions of this type can also be found in their paper.

It remains to estimate the mean bearing and the standard deviation of the distribution from each set of indicated bearings. The mean bearing can be readily obtained from the mean indicated bearing. By calculating the variance, σ^2, of the indicated-bearing distribution given in eqn. (1) it is found that

$$\sigma^2 = \frac{\nu^2}{2} \log_\epsilon \frac{4}{\nu^2} \quad . \quad . \quad . \quad . \quad . \quad (3)$$

This is shown by Whale and Delves[3]; to reach this result they had to curtail the distribution effectively at $\theta = \pm\pi/2$, but this is similar to what was done in taking the present observations. A plot of the function ν is shown in Fig. 1 with the values

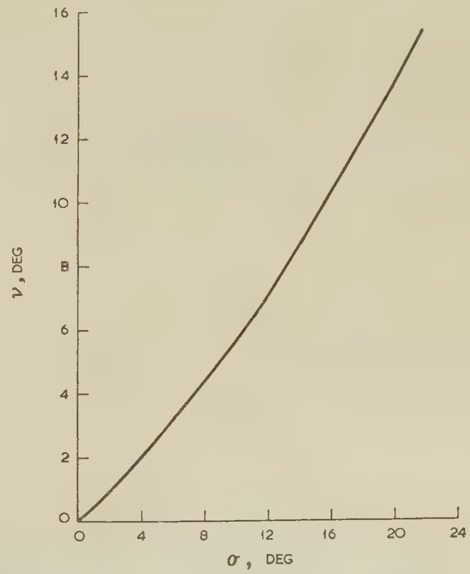

Fig. 1.—Actual standard deviation, ν, of the incident power distribution as a function of the standard deviation, σ, of the indicated bearings.

Written contributions on papers published without being read at meetings are ited for consideration with a view to publication.
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transformed from radians to degrees; the actual standard deviation is rather more than half the indicated standard deviation at values of σ in the range 5–10°, but the ratio increases towards unity as σ increases. Outside the range of values shown in the Figure an exact curve cannot be given owing to the breakdown of the approximations in the theory, but it seems reasonable to assume that the curve will continue in a similar way.

(2) RESULTS

The results given here were all obtained during the period 19th November, 1956, to 8th January, 1957. The bearings were recorded photographically on continuously-moving film, each echo recurring at a distance corresponding to the transmitter pulse-recurrence frequency of 4 pulses/sec. This was a time of high critical frequencies in the F2 layer, and consequently more than one delayed signal, as well as the forward scatter, was frequently present in the daytime. The first delayed echo arrived about 15–20 ms after the forward-scatter signal and showed a marked diurnal variation in bearing, as is illustrated in Fig. 2. The bearing of Gibraltar from Slough is 194°. The

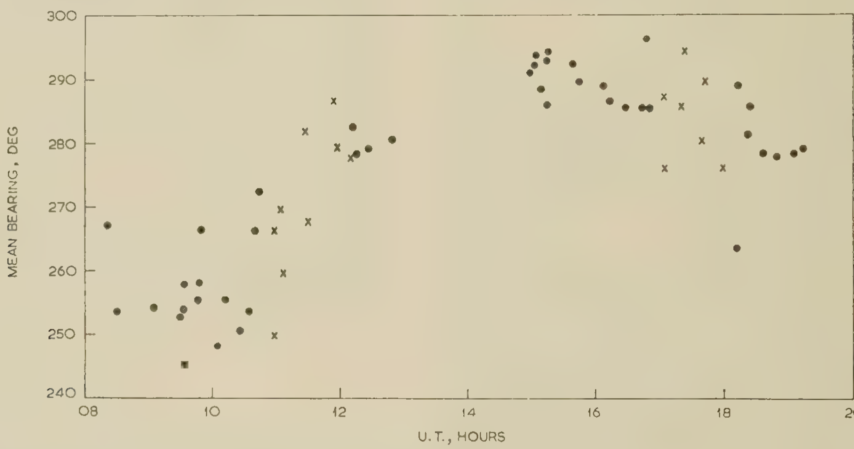


Fig. 2.—Mean bearing of the first delayed signal from Gibraltar as a function of time of day.
● Transmitter array.
× Transmitter dipole.

signal appeared at a bearing of about 250° at 0900, swung round northward to reach 290° by 1500, and moved back a little before disappearing around 1900. No observations were made between 1300 and 1500. The corresponding values of standard deviation, σ, of the indicated bearing are shown in Fig. 3. Here the trend appears to be one of gradually-decreasing standard deviation during the day. Note that the standard deviation of the actual energy distribution will be somewhat less than shown in the Figure (see Fig. 1).

It is of interest to see whether these effects can be expl theoretically. For this purpose the ionospheric predi charts for f_0F2 in December, 1956 (produced at the R Research Station) were used to calculate the area on the ea surface to each point of which propagation was possible single-hop F2 path from both Gibraltar and Slough. Thi done for the times 1000, 1200, 1600, and 1800 U.T. The cr frequencies used in the calculation were increased by a fact 1·2 as the predictions for that month turned out to be too by approximately this amount; some allowance was also for the thickness of the ionospheric layer. These charts (have been drawn on a gnomonic projection) are sho Fig. 4. A region to the south and east of Gibraltar has excluded from the attainable area in each case, as measurem of the transmitter polar diagram indicated that no radiation place in these directions. This is shown by the work of C *et al.*[2]; the polar diagram given by them applies to the a only, but they state that it is the mass of the Rock which off the south-easterly radiation. As the dipole was si beside the array it is likely that it, too, did not radiate ap ciably to the south-east.

It is evident from a consideration of the chart for 1000 the mean bearing of the area shown there is far removed from value of 255° which is found in practice. The only explana of this seems to be that the northern part of the area is r strongly illuminated than the rest, i.e. that the power radi

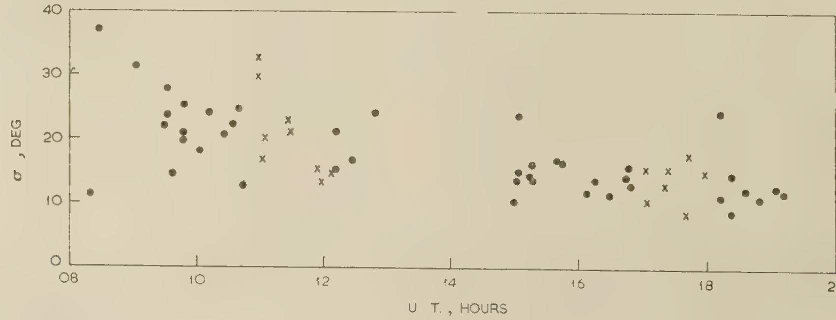
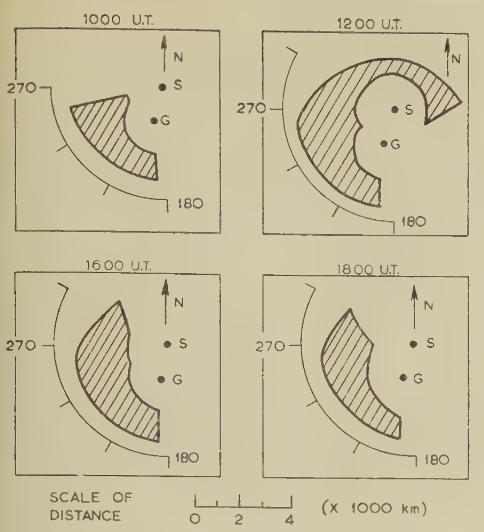


Fig. 3.—Standard deviation, σ, of the indicated bearings of the first delayed signal as a function of time of day.
● Transmitter array.
× Transmitter dipole.



. 4.—Areas (shown shaded) on the earth's surface which can be reached from both Gibraltar (G) and Slough (S) by a single reflection from the F2 layer on 37 Mc/s.

Projection, gnomonic centred on Slough.
Scales in the diagrams show bearings from Slough in degrees.
Distance scale applies only along a radial line from Slough.

m Gibraltar is greater towards the west and north-west than the south with either the transmitting array or the dipole. is hypothesis is not inconsistent with the remainder of the ults; indeed, the trend in mean bearing during the rest of the y is in agreement with this. For instance, the northerly vement of mean bearing in the morning (see Fig. 2) can be ribed to this effect combined with the spread of high electron nsities to the north as the day advances. Also the bearing 290° accompanied by a small standard deviation at 1600 is counted for by strong illumination of the area to the north-st of Gibraltar. By 1800 the mean bearing is found to have urned to about 280° as the northern boundary of the area has ved southward; although this movement is not large, it is portant because it removes a region which was previously ongly illuminated.

The second delayed signal to arrive has a delay of 25–40 ms the forward-scatter component and it too exhibits a marked rnal change of bearing, as is shown in Fig. 5. In the morning mean bearing of this signal is much more southerly than that of the first delayed echo, as is reasonable since it is only in this direction that there will be sufficiently high ionization present at distances up to 6 000 km from Slough. However, when the higher ionization spreads farther west in the afternoon, the bearing changes to values very near to that of the first delayed signal. The standard deviation, σ, varies during the day similarly to that for the earlier component, although it is slightly greater on the average.

Signals with a greater delay than 40 ms occasionally appeared on the records in a readable form. Of these, the one of the greatest interest was a signal with a very long time-delay of the order of 150 ms. Now, round-the-world signals from this transmitter have already been reported by Luscombe[4] at Slough, with a time-delay of 140 ms. The difference in the times reported are probably not significant, since in the observations discussed in this paper the nature of the display prevented time measurements of high accuracy being made.

The times of day when these long-delay signals were observed were either from 0900 to 1000 or from 1500 to 1600 U.T. A summary of the results obtained is given in Table 1. The sense

Table 1

SIGNALS WITH A LONG TIME-DELAY

Date (1956)	U.T.	Mean bearing	Standard deviation σ	Number of observations
		deg	deg	
10th December ..	0933	62	29	126
10th December ..	0948	62	24	75
10th December ..	1608	94	17	49
17th December ..	1503	127	8	26

of the bearings given in this Table cannot be taken to be definitely established; the particular values listed have been selected as the more plausible on the basis of the considerations discussed later.

To help to decide whether propagation round the world seemed possible, charts showing areas of the world in which the maximum usable frequency (m.u.f.) for the F2 layer was above 37 Mc/s have been prepared for the times in question. The boundaries of these areas have been taken to be the 30 Mc/s m.u.f. contour from the prediction charts, since, as already mentioned, the predictions for December, 1956 were too low. The charts were drawn on a azimuthal equidistant projection and are reproduced in Fig. 6.

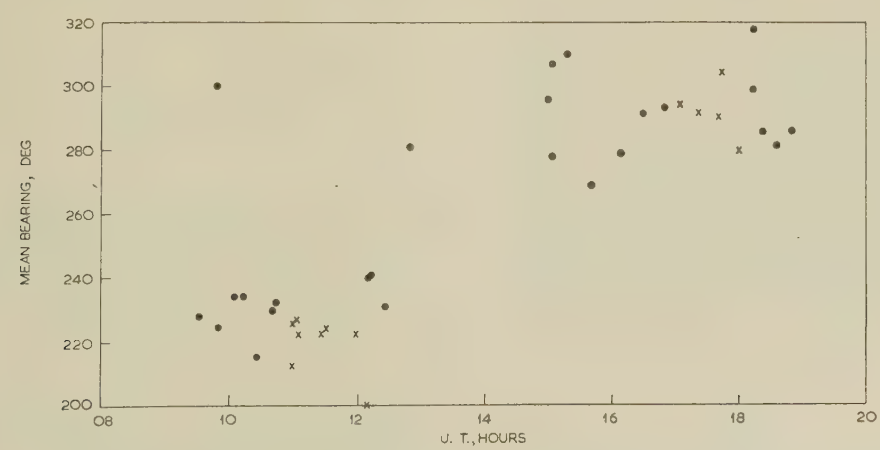


Fig. 5.—Mean bearing of the second delayed signal as a function of time of day.

● Transmitter array.
× Transmitter dipole.

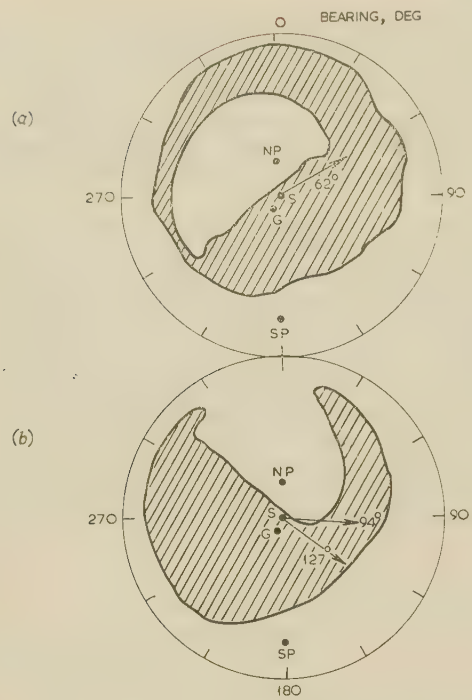


Fig. 6.—The world, with those regions shaded in which the m.u.f. in December exceeded 37 Mc/s. The measured bearings of long-delay signals are marked.

Projection, azimuthal equidistant centred on Slough. (The circumference of the circle represents the antipodes.)
S Slough. G Gibraltar. NP, SP North and South Poles.
(a) 1000 U.T.
(b) 1600 U.T.

It can be seen from Fig. 6 that propagation round the world along the great circle which passes through Gibraltar and Slough would be impossible at either time of day, and this, of course, is borne out by the results in Table 1. However, as the distance between Gibraltar and Slough is small compared with the circumference of the earth, propagation between the two would be possible along paths reaching Slough from almost any direction, as long as ionospheric electron densities permitted this. The actual path will, of course, lie quite near to a great circle passing through the mid-point M of GS. Now if in a diagram similar to Fig. 6(*a*), but centred on M, the bearing of this great circle is varied from 0 to 180°, it will be found that the length of path in regions where the m.u.f. is below 37 Mc/s falls to a fairly sharp minimum in the sector between 50° and 60°, the value of the minimum being about 22% of the total path. This agrees quite well with the observed bearing of 62°. The bearing has been chosen as 62° rather than 242° because, in going from Gibraltar to Slough round the world, the path will thereby lie clearer of the low m.u.f. region to the south-west of Slough. For the afternoon a similar calculation shows that the length of path in regions where the m.u.f. is below 37 Mc/s falls to a rather broad minimum from bearings 90° to 130°, the minimum value being just under 25% of the total path. The observed bearings of 94° and 127° agree satisfactorily with this; the 180° ambi[illegible] has been resolved in this way because the use of the opp[illegible] sense would mean that the transmitter was radiating apprec[illegible] to the east and south-east, which is contrary to the observed [illegible] diagram.

No explanation has yet been given here of how the w[illegible] traverse the regions of low m.u.f., which in both cases are n[illegible] 10 000 km across—much too long for a single hop. How[illegible] it should be remembered in the first place that the m[illegible] charts of Fig. 6 may not be very accurate as they are only b[illegible] on prediction charts. Even if they are accepted as giving ave[illegible] conditions for December, 1956, there are at least two me[illegible] nisms which would account for the propagation. One is tha[illegible] electron densities were above average when the observa[illegible] were made; the morning occurrences in particular can read[illegible] explained in this way. The other mechanism is that propose[illegible] Stein[5]; it involves reflections from tilted layers and per[illegible] propagation at frequencies higher than the m.u.f.

(3) CONCLUSIONS

In the winter of 1956–57 the signals arriving at Slough [illegible] delays of less than 40 ms relative to the forward-scatter sign[illegible] came from a mean bearing towards the west. It appears [illegible] the radiation from the transmitter is stronger in this direc[illegible] than towards the south. Signals which had travelled round [illegible] world were also observed occasionally; these had not trav[illegible] along the long great-circle path from Gibraltar to Slough bu[illegible] followed a path which traversed regions of low electron de[illegible] for only a small proportion of its length.

(4) ACKNOWLEDGMENTS

The work described was carried out as part of the progra[illegible] of the Radio Research Board. The paper is publishe[illegible] permission of the Director of Radio Research of the Depart[illegible] of Scientific and Industrial Research.

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;OME FACTORS INFLUENCING 3 cm RADIO-WAVE PROPAGATION OVERSEA WITHIN AND BEYOND THE RADIO HORIZON

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SUMMARY

ıe results of a number of short-term oversea measurements of the ıation of received signal level with range from a 10 Gc/s transmitter presented. It is shown that the variation of signal level within horizon was rarely that expected for propagation through an osphere having a uniform refractive-index gradient. Signal losses :om 5 to 30 dB frequently occurred well within the horizon, these ɜs being recovered when the range between transmitter and receiver sufficiently reduced. A well-defined interference pattern usually ırred in the region of reduced signal level. Some data on the ation of refractive index with height up to about 700 ft above sea-l were gathered using a radio sonde and a captive balloon, but detail was not sufficiently fine to enable a direct relationship to be blished between signal losses within the horizon and the occurrence rregularities in the refractive-index profile at low elevations.

. direct relationship was found to exist between the signal level ıin the horizon and that propagated well beyond the horizon into extra-diffraction region.

(1) INTRODUCTION

;ince 1949 the Admiralty has sponsored a number of short-n investigations of centimetric radio-wave propagation over paths to ranges well beyond the horizon.[1,2] This work wed that some factors affecting long-range propagation at ıuencies of about 10 Gc/s were not completely understood. particular, the signal level in the extra-diffraction region ıld frequently be some 10–15 dB lower, relative to free space, n the corresponding level at 3 Gc/s, although the rate of ıal attenuation with increasing distance at the two frequencies ıained similar. The extra-diffraction signal level and attenua-ı rate with distance at 3 Gc/s were in agreement with oospheric scatter propagation theory,[3] but this theory could explain the greatly reduced signal level experienced at Gc/s.

Vhen a new series of signal-level measurements was made ıin the radio horizon at 10 Gc/s it was found that a similar ; of 10–15 dB could occur for the more distant maxima of interference pattern. At smaller ranges, however, the signal ɛl approached the expected value, resulting in a well-defined) to the interference pattern (Fig. 1).[4] This effect was fre-ntly observed, although occasions did arise when the recovery he signal to a near free-space value was not detected, probably ause the range was not sufficiently reduced. Under these ditions it appeared that an overall loss of system sensitivity occurred. However, the use of power output monitors at transmitter, and carefully calibrated receiving equipment with lities for checking receiver sensitivity in the field, indicated t the observed effects were due to a propagation phenomenon ıer than an instrumental error.

ince only the low-angle lobes of the vertical radiation pattern the transmitter contribute to radio-wave scattering in the lower troposphere, any diversion of energy from these lobes will result in correspondingly less energy entering the volume common to the transmitter and receiver aerial beams and being scattered or reflected into the direction of the receiver. A direct relationship between the signal levels in the far interference and extra-diffraction regions would therefore be expected, so that

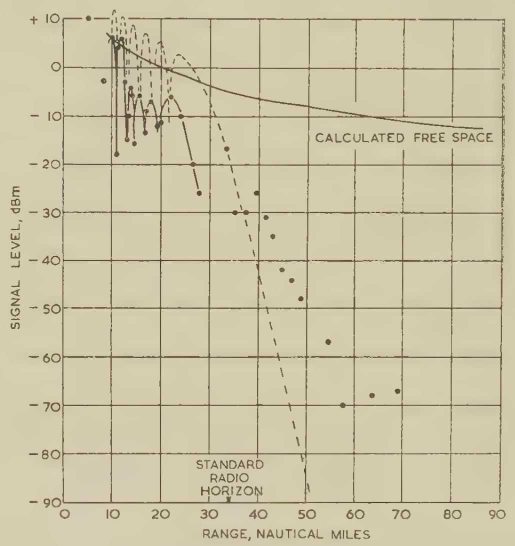


Fig. 1.—Variation of signal level with range at 10 Gc/s, English Channel, 27th November, 1957.

Transmitter aerial height 60 ft.
Receiver aerial height 400 ft.
---- Theoretical prediction for a 'standard' atmosphere.
—— Observed interference pattern.

when applying scatter theory to the extra-diffraction signal level at 10 Gc/s the measured free-space level of the contributing lobes should be used as the reference level instead of the theoretical value. When this is done there is no longer any marked discrepancy[5] in the extra-diffraction signal levels at 3 Gc/s and 10 Gc/s. It has been suggested[6] that low-level ducts are responsible for the reduced signal levels recorded within the horizon at 10 Gc/s and hence for the reduced extra-diffraction signal level commonly observed at this frequency.

An extensive radio-meteorological study would be required to establish the extent and duration of the effect of ducts on 10 Gc/s propagation over sea paths. Such a study has not been possible, but a further series of short-term measurements using both ship-borne and air-borne transmitters has been made, the results of which are now presented.

(2) EXPERIMENTAL PROCEDURE

In order to study signal-level variations within and beyond the radio horizon, transmissions from ship-borne and air-borne transmitters have been received at shore-based sites. The ship

ritten contributions on papers published without being read at meetings are ɜd for consideration with a view to publication.
e authors are at the Admiralty Surface Weapons Establishment, Portsmouth.

or aircraft carried out range runs along a predetermined track on a constant bearing from the receiving site. The opening runs started from as close to the receiving site as possible and extended to the maximum required range, the closing runs being made to the minimum range. The propagation paths were entirely over sea at all times.

The transmitters were pulse modulated and radiated from continuously rotating aerials. The shore receiver was mobile, with the aerials mounted on a rotatable cabin housing the receiving equipment. The beamwidths of the aerials used during the experiments are listed in Table 1. Horizontal polarization was used throughout.

Table 1

Aerial	Beamwidth (degrees to half-power points)	
	Horizontal	Vertical
Transmitter		
Ship-borne	1·6	30·0
Air-borne	4·0	6·0
Receiver		
6 ft diameter paraboloid	1·3	1·3
20 dB horn	10·0	9·7

The horn aerial was used for reception within the horizon and the paraboloid at greater ranges when the increased aerial gain was desired. The paraboloid was occasionally used for reception within the horizon in order to make bearing checks using a visual sight mounted on the aerial and when the use of the narrower beamwidth discriminated against interfering signals.

At the shore site the received signal level was continuously recorded as a function of time. When using a ship-borne transmitter the ship's position as a function of time was charted at sea using modern navigational aids. In the case of the air-borne transmitter the position of the aircraft was obtained from radar plots given by a 10 cm radar set installed 780 ft above the sea and overlooking the propagation path. Subsequently, the variation of signal level with time was transcribed into a signal-level variation with range from the transmitter.

(3) PROPAGATION MEASUREMENTS USING AN AIR-BORNE TRANSMITTER

When measuring the variation of received signal level with range, the use of an air-borne transmitter has the following advantages over one carried by a ship:

(*a*) Measurements of the received signal may easily be made for different transmitter heights.
(*b*) Signals may be received over a greater range of elevation angles.
(*c*) An opening or closing run takes less time to complete, so that propagation conditions along the path may not change significantly between the start and finish of a run, or between successive runs.

During April and September, 1958, an air-borne transmitter was used for studying low-level propagation over part of the English Channel. A receiver, land based on the Isle of Wight, was used at one of two sites either 50 ft or 300 ft above mean sea-level. The aircraft flew on a predetermined track at a height of 100, 250, 500 or 1 000 ft.

(3.1) Results during April, 1958

Measurements were made on five consecutive days in April, 1958, out to ranges of 100 nautical miles. Ten opening and ten closing runs were made, successive runs being at different heights.

On all days low-signal levels were received within the ho[r]
and enhanced signals were present beyond the horizon.

For measurements made on two successive days, the rece[iving] aerial was 50 ft above the sea. The variation of received s[ignal] level for transmitter heights of 100 and 250 ft is shown in Fi[gs. 2] and 3, respectively. Variations similar to those shown in F[igs.]

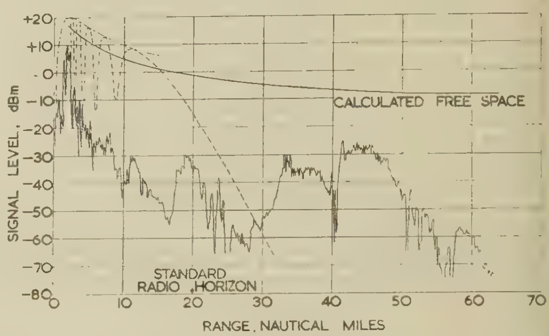


Fig. 2.—Variation of signal level with range at 10 Gc/s, En[glish] Channel, 22nd April, 1958.
Transmitter aerial height 100 ft.
Receiver aerial height 50 ft.
---- Theoretical prediction for a 'standard' atmosphere, taking account of the vertical-coverage diagrams.

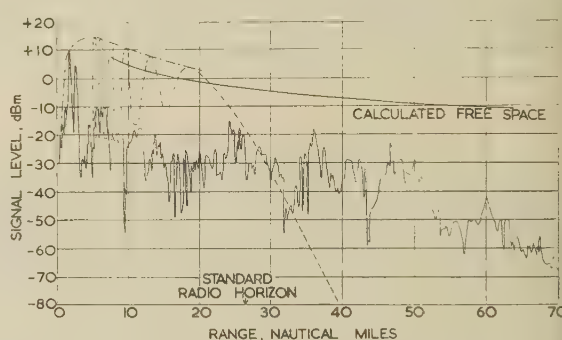


Fig. 3.—Variation of signal level with range at 10 Gc/s, En[glish] Channel, 22nd April, 1958.
Transmitter aerial height 250 ft.
Receiver aerial height 50 ft.
---- Theoretical prediction for a 'standard' atmosphere, taking account of the vertical-coverage diagrams.

occurred for transmitter heights of 500 and 1 000 ft. The [main] features of the signal-level variations recorded at this site [were:]

(*a*) The signal level approached the theoretical value at elev[ations] greater than about one degree relative to the earth tangent [at the] transmitter site.
(*b*) At smaller elevations, the signal level within the 'stan[dard'] radio-horizon range was reduced by at least 30 dB below the [free-]space level, there being no well-defined interference pattern.
(*c*) The signal level beyond the 'standard' radio-horizon [range] appeared to be influenced by the presence of a duct. Two [types] of signal variation were recorded, typified by Figs. 2 and 3.

In Fig. 2, a signal enhancement occurred beyond 30 na[utical] miles, following a severe within-horizon fade. This enh[ance]ment persisted to a range of nearly 50 nautical miles, be[yond] which the signal attenuation rate fell rapidly. It is likely [that] the signal increase centred at 40 ± 10 nautical miles was d[ue to] leakage from a duct, while beyond 50 nautical miles the s[ignal] was received by diffraction around the earth's surface.

The second type of signal variation, shown in Fig. 3, sh[owed] little change in character or attenuation rate either with[in or] beyond the horizon, suggesting the presence of a well-de[fined] surface duct.

or three successive days the receiver was 30 ft above the sea. ilar variations of signal occurred each day. During one ning a series of runs, to a maximum range of 25 nautical s, took place in succession with opening and closing runs ne same height. An opening and closing run at 100 ft was nediately followed by an opening and closing run at 250 ft, owed by similar runs at 500 and 1 000 ft. The signal varias recorded on opening runs are shown in Fig. 4. It is seen there is a signal loss on increasing the transmitter height

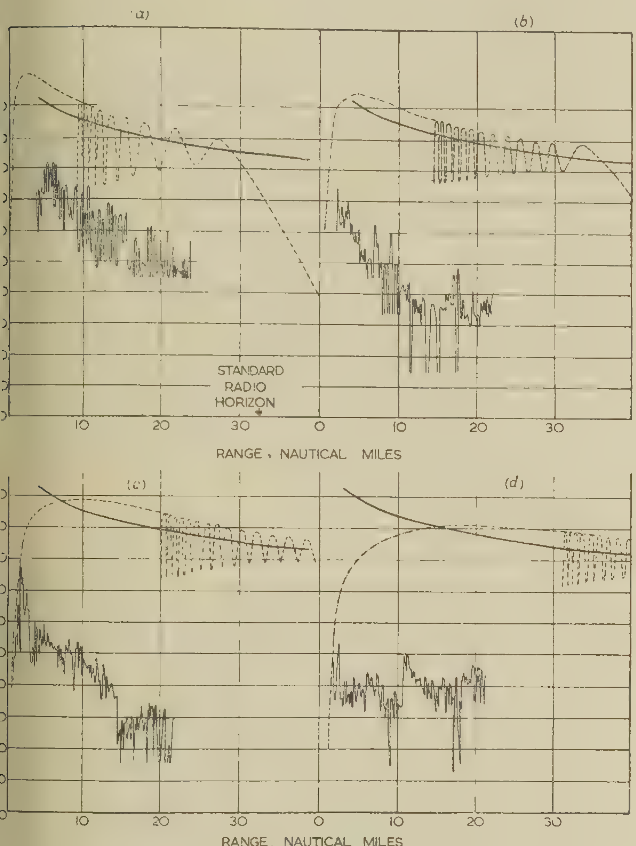


Fig. 4.—Variation of signal level with range at 10 Gc/s, English Channel, 25th April, 1958.

Transmitter aerial height (*a*) 100 ft.
(*b*) 250 ft.
(*c*) 500 ft.
(*d*) 1 000 ft.
Receiver aerial height 300 ft.
- Theoretical prediction for a 'standard' atmosphere, taking account of the aerial vertical-coverage diagrams.

n 100 to 250 ft, with no further loss on increasing the height 00 ft. In fact, a signal enhancement occurred on the 500 ft ing run between 20 and 10 nautical miles. Increasing the ht to 1 000 ft produced a further signal loss on the opening but there was no difference on the closing run from that 00 ft. These variations of signal level with height of the smitter were in addition to an overall loss of between 15 30 dB from the theoretical level for propagation through a mixed atmosphere.

(3.2) Results for September, 1958

uring September, 1958, the same type of aircraft and transer were available for two days and the same receiver sites propagation path were used as for the measurements made in April, 1958. For one day the receiver was 50 ft above the sea and opening and closing runs at the same height were made to a maximum range of 30 nautical miles. Commencing with an opening and closing run at 1 000 ft (opening run shown in Fig. 5), further runs were made successively at 500, 250 and 100 ft.

All these runs showed well-defined interference patterns with the maximum signal levels exceeding that of free space. On no occasion was a loss of signal level experienced at this site.

The following day the receiver was in position 300 ft above sea-level. A similar series of opening and closing runs was performed, but the interference pattern was broken up and ill-defined, with the maximum signal level barely reaching the free-space level when the transmitter was at 100 ft. On increasing the transmitter height to 250 ft, an overall drop of level of about 5 dB occurred, this loss persisting up to 1 000 ft.

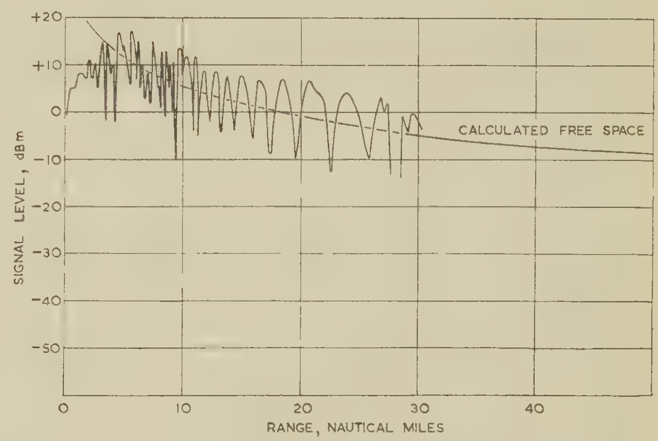


Fig. 5.—Variation of signal level with range at 10 Gc/s, English Channel, 15th September, 1958.

Transmitter aerial height 1 000 ft.
Receiver aerial height 50 ft.
Opening run duration: 1011–1023 B.S.T.

(4) PROPAGATION MEASUREMENTS USING SHIP-BORNE TRANSMITTERS

Measurements of received signal level using a 10 Gc/s ship-borne transmitter were made in the English Channel during September, 1958, and July, 1959. During September, 1958, the same receiving sites were used as in the series of experiments using the air-borne transmitter, again using one receiver. During July, 1959, two receivers were available, making possible simultaneous reception at two heights.

(4.1) Results for September, 1958

Measurements were made over a period of two weeks, the first with the receiver at 50 ft above mean sea-level and the following week with the receiver at 300 ft. The ship, with the transmitter aerial 60 ft above the sea, made opening and closing runs with maximum ranges varying between 25 and 60 nautical miles. The main features of the received signal in the interference, diffraction and extra-diffraction regions are summarized in the following Section.

(4.2) Receiver at 50 ft above Sea-Level

(4.2.1) Interference Region.

(*a*) Of a total of four opening and four closing runs, five showed a loss of signal within the horizon while the remaining three showed no significant loss and had maximum signal levels greater than free space.

(b) When a loss of signal level occurred, the free-space level was approached or exceeded when the range was reduced to about 1 or 2 nautical miles from the transmitter.

(c) A well-defined interference pattern was usually recorded, even in the regions of reduced signal level. The spacing of adjacent maxima and the depth of successive minima were not predictable on the assumption of propagation through an atmosphere having a uniform variation of refractive index with height, even when the maximum signal levels exceeded the free-space level.

(d) On one occasion at a range of about 8 nautical miles, a signal enhancement of 10 dB above the maximum expected level for two-path propagation was observed. A similar enhancement was recorded during the series of measurements using the air-borne transmitter during September, 1958. It seems likely that multi-path propagation was responsible for these signal increases.

(4.2.2) Diffraction Region.

(a) The signal attenuation rate in the diffraction region appeared uncorrelated with the degree of signal loss that occurred in the interference region.

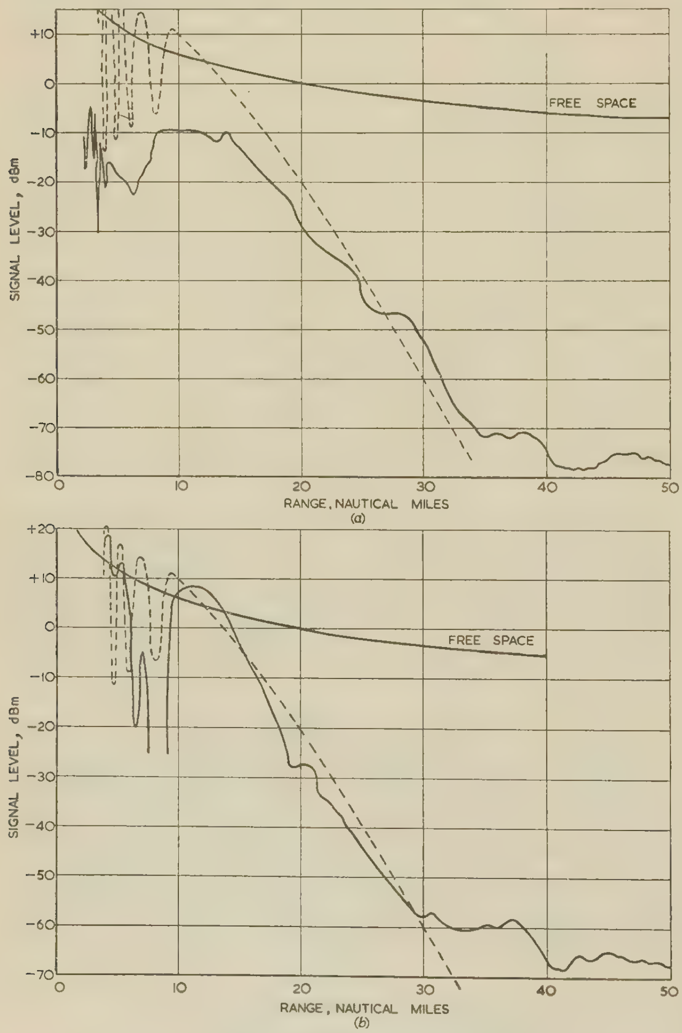


Fig. 6.—Variation of signal level with range at 10 Gc/s, English Channel, 1st September, 1958.
Transmitter height 60 ft.
Receiver height 50 ft.
(a) Opening run.
(b) Closing run.
---- Theoretical prediction for a 'standard' atmosphere.

(b) The attenuation rate was either similar to, or rath than, expected for propagation through a 'standard' atmos This suggests that, beyond the horizon, the relevant part atmosphere important for propagation had a refractive gradient which was either near the 'standard' value or one tended to produce slight super-refraction.

(c) When the interference pattern suffered an overall d level, the signal in the diffraction region either suffered a drop or regained a value near that expected after propa through a 'standard' atmosphere.

(4.2.3) Extra-Diffraction Region.

Fig. 6 shows range runs made on the same day to ran which the extra-diffraction signal could be recorde Fig. 6(a) the signal level within the horizon is some 15 below the free-space level and the extra-diffraction signal 65 dB below. Fig. 6(b) shows an increased signal with horizon, with the interference maxima near the theoretical and an extra-diffraction signal level about 55 dB belo free-space level.

Measurements of extra-diffraction signal levels at 3 Gc/s on many occasions over a period of several years,[1] have that at the threshold of the extra-diffraction region the level is about 50 dB below the free-space level. Most m ments at 10 Gc/s have indicated signal levels about 65 d the free-space level at the threshold.[2] Figs. 6(a) and (b) appear to confirm that there is a direct relationship betw signal level at ranges just within the radio horizon and the extra-diffraction region.

(4.3) Receiver at 300 ft above Sea Level

The receiver was at 300 ft for five consecutive day opening and a closing run was made each day to a ma range of between 30 and 50 nautical miles. The types of variation recorded at this site may be summarized as foll

(a) Signals up to or exceeding the free-space level were at ranges less than 2 to 4 nautical miles, followed by a 'step interference pattern. This was well defined and the maxima from 5 to 20 dB below the free-space level. These effects we on five of the ten possible occasions.

(b) The signal was influenced on one occasion by super-ref The maxima of the interference pattern approached or excee free-space level at all ranges, and the diffraction attenuation less than would be expected for propagation through a 'st atmosphere.

(c) The signal was influenced on one day by well-defin propagation, and the average attenuation rate appeared unc at all ranges out to 50 nautical miles.

(5) SIMULTANEOUS MEASUREMENTS AT TWO AE HEIGHTS

During July, 1959, range runs were made in the E Channel using two receivers, thus making possible simult measurements at two aerial heights. The aerial of th borne transmitter was 50 ft above the sea. At high t receiving aerials were 10 ft and 40 ft above sea-level, and tide, 22 ft and 52 ft above sea-level. Measurements were over a period of two weeks, 28 opening and closing run made to a maximum range of 15 nautical miles. The mi ranges were limited by the state of the tide. At high t ship was able to close the range to about 1 nautical mile.

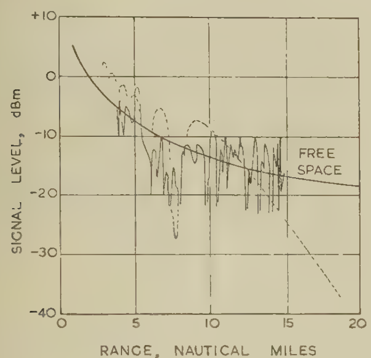
During the time that measurements were made there w change in the propagation conditions, and the signal vari recorded at the two aerials may be summarized as follow

(a) The low aerial was always influenced by the presence level ducts along the propagation path, a near free-space being maintained to the maximum range involved. At th

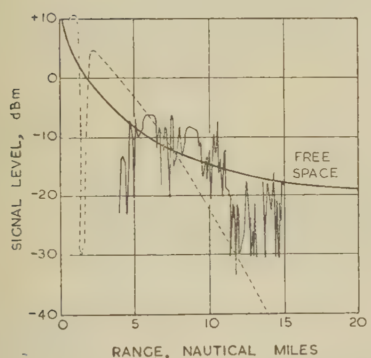
signal was then some 15 dB greater than expected for propaga-ı through a 'standard' atmosphere [Figs. 7(*b*) and 8(*b*)].

5) The high aerial was influenced by super-refraction for about ı-thirds of the time that measurements were made, giving a uced attenuation rate at ranges beyond that of the last inter-ınce maximum [Fig. 8(*a*)]. Under these conditions the spacings the maxima and minima of the interference pattern suggested pagation over an earth of effective radius about twice the metrical radius. For the remaining third of the time the inter-ınce pattern was broken up and the received signal was of the e expected under conditions of ducting or multi-path propagation ş. 7(*a*)]. Only on one occasion was the level of an interference kimum significantly below the free-space value (by about 10 dB). ıerwise there was no loss of signal within the horizon, the maxi-m levels being at, or above, the free-space level at all times.

appears that during the time that this set of measurements nade the refractive-index structure of the atmosphere at low ; along the propagation path favoured super-refraction at ıts above the normal surface duct.



(*a*)



(*b*)

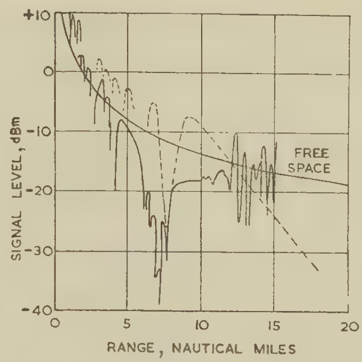
Fig. 7.—Variation of signal level with range at 10 Gc/s, English Channel, July, 1959.

Transmitter height 50 ft.
Receiver height: (*a*) High aerial, 52 ft.
(*b*) Low aerial, 22 ft.
---- Theoretical prediction for a 'standard' atmosphere.

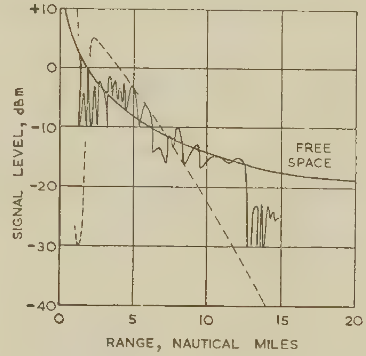
SUMMARY OF SIGNAL TYPES RECORDED WITHIN THE HORIZON

ıree main types of signal variation with range from a 10 Gc/s smitter have been recorded in the present series of tigations.

Type A.—The signal level at all ranges less than that of the last erference maximum is as expected for propagation through an nosphere having a uniform variation of refractive index with ght. The maxima of the interference pattern are then nearly B above the free-space level. At ranges greater than that of the t interference maximum the effects of a surface duct usually come apparent because of the smaller propagation angles involved.



(*a*)



(*b*)

Fig. 8.—Variation of signal level with range at 10 Gc/s, English Channel, July, 1959.

Transmitter height 50 ft.
Receiver height: (*a*) High aerial, 40 ft.
(*b*) Low aerial, 10 ft.
---- Theoretical prediction for a 'standard' atmosphere.

Type B.—The signal has a maximum value above the free-space level to ranges between 2 and 5 nautical miles. At greater ranges a drop in overall level of from 5 to 30 dB may occur, usually with the interference pattern remaining well defined. Beyond the diffraction region the weak extra-diffraction field is recorded.

Type C.—The signal within the horizon is similar to type B, but there is no well-defined diffraction region, the signal apparently being propagated within the boundaries of a surface or elevated duct to ranges well beyond the normal radio horizon.

Since the measurements that have so far been made were over short-term periods, it is not possible to give detailed statistics of occurrence of the various types of signal within the horizon over any particular propagation path, although the following general statements can be made:

Type A signals have occurred rarely, type B being predominant during the series of measurements. Type C has been recorded mostly when using air-borne transmitters, when the presence of elevated ducts becomes important in influencing signal levels within and beyond the horizon.

(7) POSSIBLE EXPLANATION OF OBSERVED EFFECTS

In temperate latitudes a well-mixed or so-called 'standard' atmosphere has a modified refractive index, M, which increases with height at a more or less constant rate of 0·036 M units/ft. The modified refractive index is defined as

$$M = \left(n - 1 + \frac{h}{a}\right) \times 10^6$$

where n is the refractive index of the atmosphere at height h above the earth's surface and a is the radius of the earth.

Calculations of field strength within the radio horizon are usually based on ray theory with the assumption of a constant value for dM/dh from ground level to heights well above the propagation path. This permits the assumption of rectilinear propagation over an earth of modified radius, the 'radio' earth radius being larger than the geometrical one under conditions of super-refraction and smaller under sub-refractive conditions. The resulting variation of field strength with range within the horizon then consists of a number of maxima and minima resulting from the vector addition of direct and sea-reflected rays. If the sea was a perfect reflector, the reflected rays would be reduced in amplitude only by divergence on reflection from a spherical surface and the interference maxima would be nearly 6 dB above the free-space level at that range. In practice, diffuse reflections and varying reflection coefficients reduce the maximum signal level below that calculated, but in general the interference maxima would be expected to exceed the free-space level.

Variations of temperature and water-vapour lapse rates in the atmosphere produce gradients of M that persist in limited height intervals. The resulting M/h profile then shows a number of maxima and minima of M, and rectilinear propagation over an earth of modified radius can no longer be considered when these variations occur within the propagation path. The theoretical treatment of propagation under these conditions is most effectively performed by ray-tracing methods using the actual M/h profile, supplemented, where required, by wave theory. Predictions of signal level from ray tracings are made by comparing the ray densities in the 'standard' atmosphere with those for the actual atmosphere involved. Within a caustic system, where the ray density becomes infinite, numerical values of the signal level can be obtained only from wave theory. Methods of ray tracing using an analogue computer have been used successfully by Wong[7] in an investigation of within-horizon fade-outs experienced during air-to-air and ground-to-air propagation at metric and centimetric wavelengths. A direct relationship was found between the magnitude and extent of fade-outs, or 'radio holes', experienced at 3 295 Mc/s, and measured refractive-index profiles that were obtained near the mid-point of the propagation path. Similar methods were used by Hay and Poaps[8] to study qualitatively the cause of signal fade-out within the horizon at 2 Gc/s.

The lack of adequate data on air temperature and humidity at sufficient intervals along the propagation path has so far prevented any correlation between the observed signal-level variations at 10 Gc/s and experimental refractive-index profiles. Direct measurements of air temperature and humidity, using a radio sonde suspended from a captive kite balloon, were made during September, 1958, and July, 1959, when ship-borne transmitters were used. The captive balloon was flown from the transmitting ship in September, 1958. However, during July, 1959, a second ship was used to control the balloon, and sonde ascents were made on this occasion within the propagation path instead of at one terminal. Soundings made within the path are likely to yield more representative data on the refractive-index structure affecting the propagation than soundings made at one terminal, although ideally many simultaneous soundings should be made at intervals along the path in order to build up a complete picture of the index structure, both in height and extent. The received signal is a function of the refractive-index structure along the whole path, and unless the lower levels of the atmosphere are horizontally statified, one sounding alone will not enable theoretical predictions of signal level to be made using ray-tracing techniques. In general, the soundings did show the presence of surface ducts which affected low-angle propagation when type A signals were recorded, but on no occasion did the soundings show sufficient detail to account for the severe within-horizon effects that are characteristic of
signals.

During the period in April, 1958, when measuremen
made of the received signal from an air-borne transmitter
layer about 600 ft thick persisted over the sea and the ob
trapping into horizontal layers of smoke from ships
across the propagation path indicated that the air over
was very stable. It is during these conditions that trap
electromagnetic energy will almost certainly occur.

The presence of super-refractive or sub-refractive laye
the height of the transmitter aerial can cause a marked red
tion of near horizontal energy from the transmitter.[7]
layer is of sufficient intensity to provide a negative ref
index gradient, trapping of the energy may occur within
duct, the duct becoming more effective as the frequency
radiation increases. If the transmitter is just above
having a negative gradient, radiation entering the layer a
angles of incidence will be refracted downwards, the am
refraction becoming less as the angle of incidence becom
This situation is shown in Fig. 9(a), which represents
tracing drawn for a flat earth, the ray curvature being a
accordingly. The variation of modified refractive index t
the layer is shown by the dashed curve at the left of
tracing. Above the layer the energy distribution is unaff
the presence of the layer and the field strength may be cal
using the value of dM/dh above the layer. Below the bo
TA the field has been modified owing to the effect of th

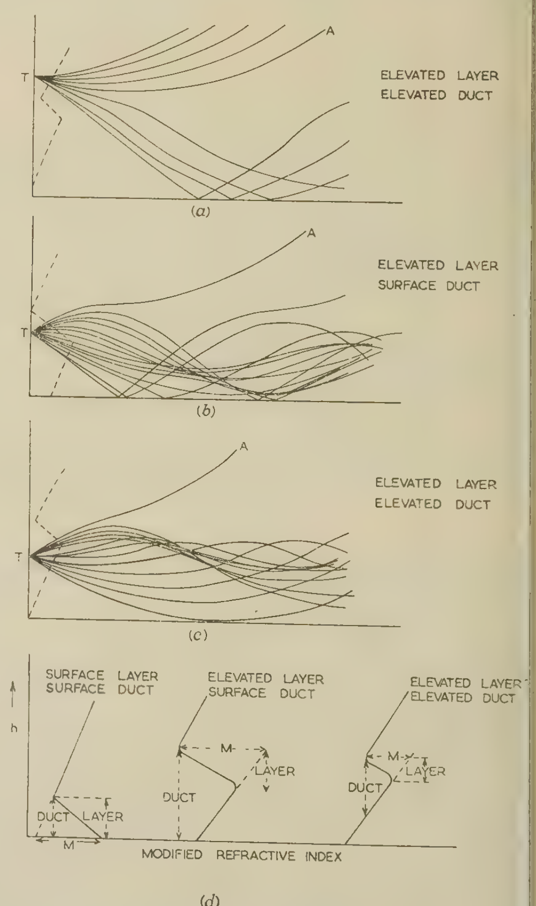


Fig. 9.—Ray tracings.

(a) Transmitter above an elevated layer.
(b) Transmitter within an elevated layer.
(c) Transmitter below an elevated layer.
(d) Distinctions between surface and elevated layers and surface and eleva

·acings for the transmitter within and below the layer are ι in Figs. 9(*b*) and (*c*), respectively. Fig. 9(*d*) illustrates stinction between the terms surface and elevated ducts and e and elevated layers used to classify Figs. 9(*a*), (*b*) and (*c*). sudden change of field strength which produces the 'step' near-interference-region characteristic of many records of -level variation with range at 10 Gc/s would occur across oundary TA, while the subsequent variation of received trength with increasing range would depend on the varia-f layer height and intensity with range and on the height receiving terminal.

(8) CONCLUSIONS

ːrsea measurements of received signal level as a function of from a 10 Gc/s transmitter have shown that at ranges well ι the radio horizon the level of the interference maxima can ently be from 5 to 30 dB below the level expected for gation through an atmosphere having a uniform variation ractive index with height. On the occasions that this loss nal level occurs, an increase in level to a near theoretical takes place when the range between the transmitter and er is sufficiently reduced. This increase occurs at the where it is considered that propagation is no longer nced by discontinuities in the refractive-index structure of tmosphere close to the sea surface because of the larger ıgation angles involved (in excess of one degree in elevation). present the data gathered on the variation of refractive up to heights of about 700 ft during times that signal-level ırements were made has not provided sufficient detail to it a correlation between the existence of trapping layers he occurrence of reduced signal levels within the horizon. rolonged radio-meteorological investigation would be red to study this and the statistics of occurrence and extent ch layers. A common feature of all the measurements has that the receiving terminal has always been on shore, and it is possible that the effects observed are due to the coastal environment. It might be expected, therefore, that the type of weather, wind direction or type of coastline would be significant. In fact, no apparent influence of these factors has been detected.

(9) ACKNOWLEDGMENT

The paper is published by permission of the Admiralty.

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PROBABILITIES OF INTERFERENCE WITH MOBILE FIELD RADIO DERIVI FROM A FIELD-STRENGTH SURVEY AT 59 Mc/s

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(*The paper was first received* 21*st October*, 1959, *in revised form* 5*th July, and in final form* 11*th October*, 1960.)

SUMMARY

The paper gives an account of an experimental investigation of common- and immediately-adjacent-channel interference areas for mobile radio networks, based on a field-strength survey at 59 Mc/s over flat and hilly country in north-west Germany. The variation of field strength with distance is shown to be in close accord with calculation. Particular attention has been paid to investigating the distribution of field strength at each distance, which is shown to approximate closely to log-normal. It is also established that the variance of these log-normal distributions may be considered to have a single value for each type of country, irrespective of range. Protection ratios acceptable for satisfactory operation are determined from experimental observation of common- and adjacent-channel interference. The derived data are used to determine probabilities of interference and are presented graphically for practical use.

LIST OF SYMBOLS

A = Wanted transmitter on link AB.
B = Receiver on link AB.
l = Distance from A to B.
C = Interfering transmitter.
x = Common-channel interference distance at which C, on the same frequency as A and B, just precludes satisfactory reception of A at B.
H_1 = Aerial factor of measuring set.
I_c = Common-channel protection ratio, the wanted station being just clearly intelligible; i.e. (wanted-station field strength)/(interfering-station field strength), dB.
I_{1A} = Immediately-adjacent-channel protection ratio, the wanted station being just clearly intelligible, i.e. (wanted-station field strength on wanted-station frequency)/(interfering-station field strength on interferer's frequency), dB.
s^2 = Variance.
s = Standard deviation.
s_F, s_H = Standard deviation of log field strength for flat and hilly country, respectively.
E_0 = Field strength over spherical smooth earth, volts per metre.
P = Power radiated, watts.
G = Gain of radiator, with respect to $\frac{1}{2}\lambda$ dipole.
h_T, h_R = Heights of transmitting and receiving aerials respectively, m.
d = Distance, m.
E_f = Median field strength over irregular terrain.
e_l = $\log_{10}$ field strength likely to be received at B from wanted station A at distance l.
e_x = $\log_{10}$ field strength likely to be received at B from interferer C at distance x.
$\bar{e}_l, \bar{e}_x$ = Mean values of e_l and e_x.
$u = e_l - e_x$
$\bar{u}$ = Mean value of u.
i = Acceptable protection ratio, expressed as a log to the base 10.
s_u = Standard deviation of u.
s_l, s_x = Standard deviations of likely values of e_l respectively.
X = Units of standard deviation.
E_{fl}, E_{fx} = Median field strengths at distances l and x tively, decibels relative to 1 μV/m.
$N = E_{fl} - E_{fx}$.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Major Thomas is with the 28th Signal Regiment.

(1) INTRODUCTION

The majority of mobile v.h.f. communication systems in the congested frequency band 30–300 Mc/s.[1] In Euro allocation of frequencies below 100 Mc/s is often diffic police, fire and ambulance services, taxis and the operate in this part of the band.[2–6] To meet all dema communications within a given area it is often neces repeat frequencies and use adjacent channels with the mi spatial separation. The paper is concerned with investi to determine the minimum acceptable distances betwee operating on common and immediately adjacent chann a reasonable probability of freedom from interference. range mutual-interference effects, although important number of stations operate in the same proximity, such as or adjacent vehicles, are not considered in the paper.[5]

The investigations were primarily concerned with rad designed for mobile use in the v.h.f. band employing fre modulation and providing a simplex speech circuit us same frequency for transmission and reception. For th of set powers vary between 20 and 30 watts and a frac a watt according to the role a set is designed to fulfil. directional vertical rod aerials are normally used. arrangements for reading a weak signal in the presen stronger one are not in general use and are therefo considered.[7, 8]

The common-channel interference distance may be de that distance, x, at which transmitter C, on the same fre as transmitter A on a link AB of length l, causes just su interference at receiver B to preclude satisfactory re of the wanted transmitter A (Fig. 1). The interpreta 'satisfactory reception' depends on the type of service r and could vary from just being able to pass a message presence of interference to complete freedom from inter Whether interference occurs at B depends on the ratio of to interfering signals received. For each type of set th protection ratio, i.e. the ratio of the wanted to int signal levels, at which interference is just acceptable. Th is not always easy to determine, since for voice wor depends on the operator, as well as the type of modulat the grade of service required.[5] Owing to the 'capture' e f.m. working the ability of the operator to read throug ference is not so important as it is with amplitude m tion.[5, 6, 9] Acceptable protection ratios for the sets use been determined experimentally.

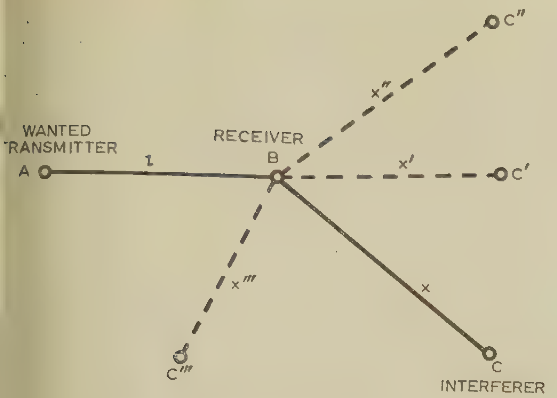
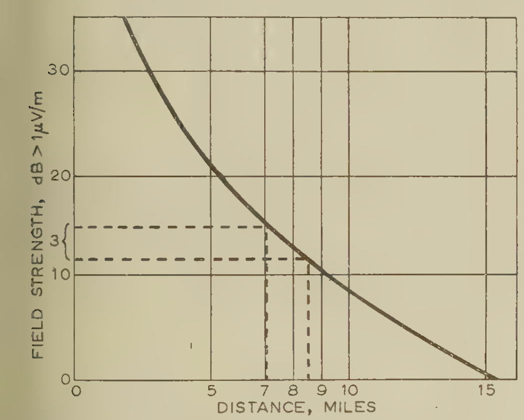


Fig. 1.—Interference distance, x.

he desired protection ratio is known and the received levels at B can be estimated, it is possible to say whether kely to interfere with the reception of A at B. For example, . 2, if the protection ratio is 3 dB, A, B and C are deployed



.—Estimation of likelihood of common-channel interference om field-strength curve at 59 Mc/s for 15-watt set with 8 ft-rod erial over good ground.

$\epsilon = 15$
$\rho = 10^{-2}$ mho/m

own in Fig. 1, all operating on 59 Mc/s, $l = 7$ miles and) miles, interference will probably not occur, whereas if miles and $x = 8$ miles, interference will probably occur. readily be appreciated that different positions of C, at C′, nd C″′, different distances, x', x'' and x''', from B could produce the same signal strengths at B, depending on the ence in the paths due to obstacles such as hills, buildings or s. The probability that interference will or will not occur e estimated if the likely variations of field strength over ent transmission paths of the required distances are known. could be determined from an analysis of observed field gths at various distances over different types of terrain, ded that there was a sufficient sample at each distance. graph of field strength against distance shown in Fig. 2 is lated for a set with 15 watts output into a vehicle-mounted al rod aerial (gain 2 dB) from standard propagation curves.[10] her it is permissible to use such curves can be determined rrying out a field-strength survey over the types of country considered and making a comparison. If the curves are used only for the determination of interference distances it necessary to define field strength in absolute units, although ould be desirable. The important point is that the slopes calculated and measured curves relating field strength to distance should be the same.[9] In the paper only normal propagation is considered and no account is taken of special propagation mechanisms.[11–14] Above 30 Mc/s normal propagation depends largely on the configuration of the terrain, including obstacles, and to a lesser extent on ambient conditions.[15–17] An investigation has therefore been undertaken to determine the propagation characteristics at 59 Mc/s over different transmission paths varying between 1 and 21 miles, and a few measurements have also been made at 38 Mc/s. The results are presented in a form suitable for the quick and easy determination of common- and adjacent-channel-interference areas in the field.

Somewhat similar work has been carried out in the United States and Britain for television and mobile-radio service areas.[9, 16, 18] It is thought, however, that this is the first work to show that field strengths from nearly randomly sited transmitters, at a given distance, approximate closely to a log-normal distribution with a variance dependent on the type of terrain irrespective of range.

(2) EXPERIMENTAL EQUIPMENT AND PROCEDURE

(2.1) Transmitters and Receivers

Two types of set were available for the field trials, namely a modern high-quality v.h.f. transportable set of 15 watts nominal output and a portable v.h.f. set with an output of less than 1 watt. The field-strength survey obviously had to be based on the more powerful set, and even this output was barely adequate, severely restricting the range of the trials. In other respects this set was well suited to the task, since it was robust, easy to operate, had accurate tuning and a power pack delivering almost constant input. Sets were installed in 4-wheel-drive vehicles to provide standardized mobile radio terminals. Two types of aerial were used—a vehicle-mounted rod and an elevated aerial, both radiating vertically polarized waves. The manpack set was only used in the determination of interference levels.

(2.2) Measurement

The field-intensity meter available was capable of measuring accurately only above 10 μV/m, giving a quite unacceptable range limitation, and thus a trials-set receiver was adapted for use as a measuring set. The meter was calibrated directly in microvolts per metre, while the receiver calibration graph showed the variation of field strength in microvolts per metre times H_1 with limiter grid current. This graph gave ambiguous readings for inputs greater than about 50 μV/m, and therefore the measuring set could not be used for the higher field strengths. The factor H_1 is associated with the elevated aerial used with the measuring set. Since this factor could not be calculated accurately, although estimated to be between 3 and 5 dB, field-strength observations made with this set were recorded in microvolts per metre times H_1. The field-strength-survey results were later used to obtain a more accurate estimate of H_1

(2.3) Determination of Protection Ratios

Protection ratios for satisfactory operation of the vehicle and manpack sets were determined experimentally, initially over comparatively short distances on an emergency airfield. The method was to move two transmitters with respect to a stationary receiver until the wanted station was just clearly intelligible through the interference. When this position—which was quite critical in terms of distance—was found, the field strengths from the transmitters were measured at the receiver. The ratio between the wanted and interfering field strengths was taken as the acceptable protection ratio. The immediately-adjacent-channel protection ratios were determined in a similar manner.

In view of difficulties associated with field-strength measurements at the longer ranges during the early stages of the survey, a system of common-channel voice-interference tests was introduced to provide an alternative means of estimating probabilities of interference. On occasions these tests gave results where the wanted station just remained a good working circuit. This provided a valuable check of protection ratios when the measuring set became available, because the interference could be observed aurally and the field strengths measured using the same instrument and aerial.

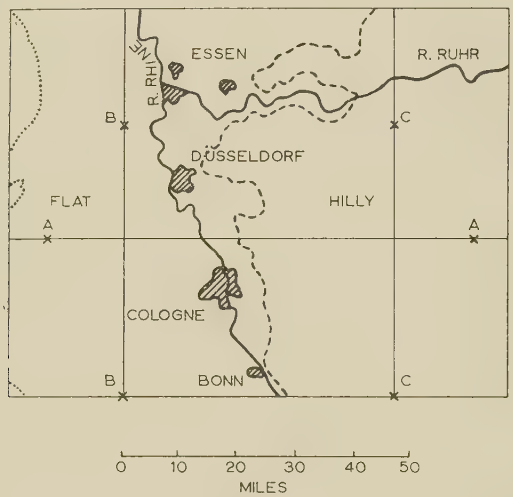


Fig. 3.—Area of field strength survey showing boundary between flat and hilly country.

(2.4) Field-Strength-Survey Procedure

The setting for the trials was north-west Germany, study of the map shows that, although there are many types of country, the majority can be described as

Flat.—Flat open or gently rolling country with gentle g and contour differences between 50 and 300 ft.

Hilly.—Broken country with alternating hills and stee valleys, in general with steep gradients and contour differ between 250 and 1 500 ft.

Both types of country include large woods and built-u The areas selected are shown at Fig. 3. Fig. 4 gives cross-sections through the flat and hilly country surve area of nearly 4 000 square miles stretching from n Dutch border to well into the Sauerland, roughly bou the River Ruhr in the north and a line east–west south The main trials took place between September and Ma weather conditions from warm sunshine to snow and absence of leaves on deciduous trees during the winter is unlikely to be significant, since conifers are prepond the woods of this country.[18]

Although it was desirable to make observations at lo mediate and high frequencies in the band, this was not p The main survey was carried out at 59 Mc/s with a sma survey at 38 Mc/s; 59 Mc/s was satisfactory through trials, but operation on 38 Mc/s was always difficult, o outside interference.

Within each type of country twelve receiver sites were and from each of these four routes following suitable r tracks were mapped out. Along each route the mile were plotted for radial distances of 1–21 miles. At each site the control station was established, complete with m apparatus. The mobile stations proceeded along the

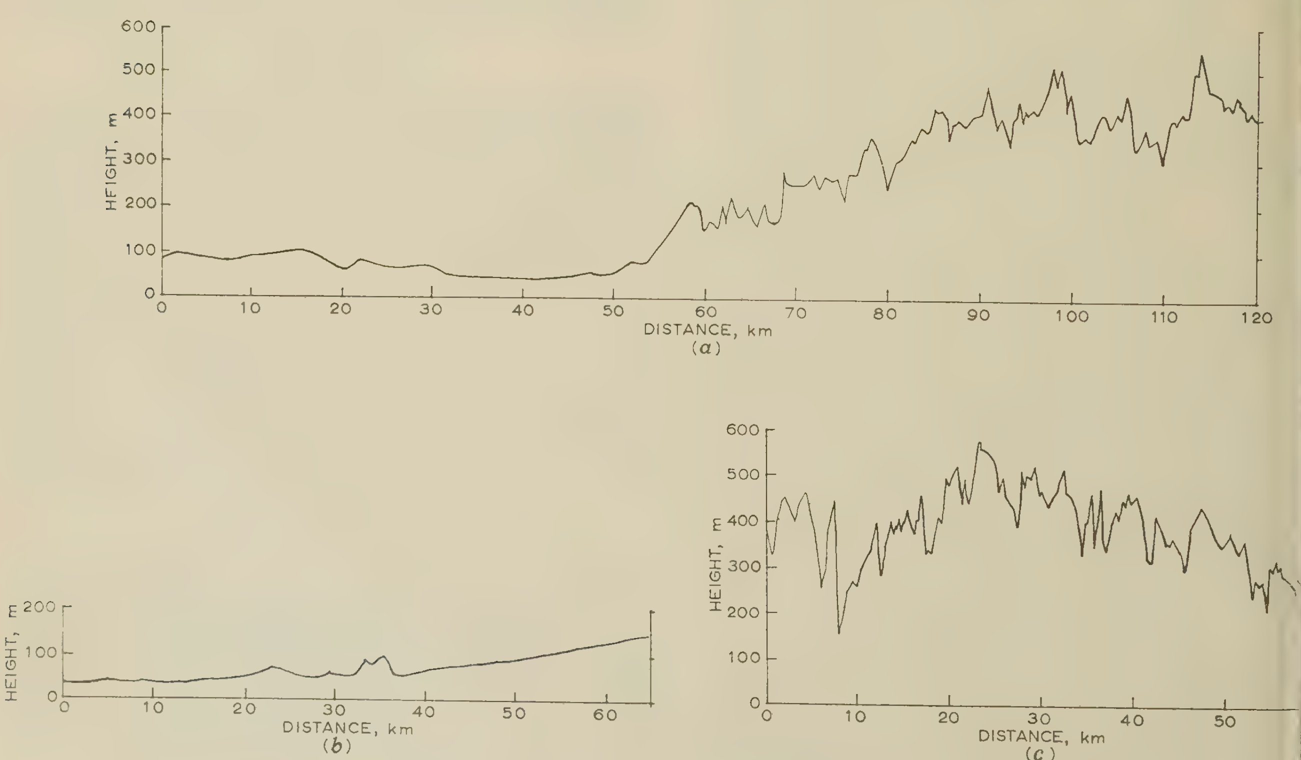


Fig. 4.—Sample contour profiles of field-strength survey area.

(*a*) West–east along A–A in Fig. 3.
(*b*) North–south along B–B in Fig. 3.
(*c*) North–south along C–C in Fig. 3.

, stopping at each mile point and transmitting. Control ved, measured and recorded the field strength received each transmission, and voice-interference tests were carried hen the outstations reached suitable positions along their . Transmissions were made using the 8 ft vertical rod and the elevated aerial, although the latter was only used imited number of positions, because of the time taken to the mast.

an unbiased survey all positions should be chosen at m. Mobile radio sets, however, are not normally sited mly: within a general area determined by the task a site ormally be chosen to provide good v.h.f. communication he other stations on the net or link with which the set is ated. The conditions when considering possible inter-e are therefore a receiver and transmitter on the same link t, sited to provide good communications with each other n interfering transmitter sited to give good communications ut regard to the interfered-with receiver. One of the cteristics of communication in the v.h.f. band is that a change in the position of the receiving aerial may produce siderable change in the received signal,[16] and this is usually into account by the operator when establishing com-cation. Since the success of a field-strength survey depends sults, very unfavourable sites should be rejected, although nust be taken not to choose receiver sites which dominate irrounding countryside. Sites were chosen with these con-itions in mind, but map references for receiver-site areas chosen at random. A reconnaissance was then made and ctual site selected to give reasonable v.h.f. communication as wide an arc as possible. It was impracticable to recon- the transmitter sites, in view of the number involved, and a certain latitude in final choice of site was allowed. On l at the map reference location the mobile station was at a reasonable position within 100 yd of this to give good und v.h.f. communication. If control could not be con-l the mobile station was permitted to move up to 50 yd to lish communication. These siting rules approximate to al field practice, except that the interferer is sited favourably respect to the receiver more often than would be expected. rrors in radial distance due to the discretion allowed in hoice of outstation sites are certainly no more than those ted in the general planning of mobile radio systems and ot considered to be significant. For each type of country, ugh a maximum of nearly 1 000 mobile-station sites are dered, only 12 base sites are used. It is estimated that any hus introduced is not significant in that the base receiver were carefully selected to be representative average field and not sited with reference to the mobile-station positions.

(3) ANALYSIS OF RESULTS

(3.1) Protection Ratios

experiments to determine acceptable protection ratios carried out on an emergency airfield at ranges between 300 300 yd. The highest and lowest values of I_c recorded 32 observations during the controlled tests were 4·0 and 3 respectively. This spread of results is probably partly the three sets not being on exactly the same frequency at ne of the observations, in spite of the precautions taken.

m the observations of voice-interference tests during the trength survey 12 instances were recorded where the on-channel interference approached the limiting case. ighest and lowest values of I_c recorded were 1·7 and 5·8 dB. r tests were carried out with the manpack set and the s are given in Table 1.

Table 1

PROTECTION RATIOS

Set	Controlled tests		Field trials	Weighted mean
	I_c	I_{1A}	I_c	I_c
	dB	dB	dB	dB
15-watt vehicle	2·8	−36	3·2	2·9
Fractional-watt manpack	3·5	−38	—	3·5

(3.2) Voice Interference Tests

In all, some 120 observations of interference were made during the survey, and a record of these is shown in Fig. 5; the

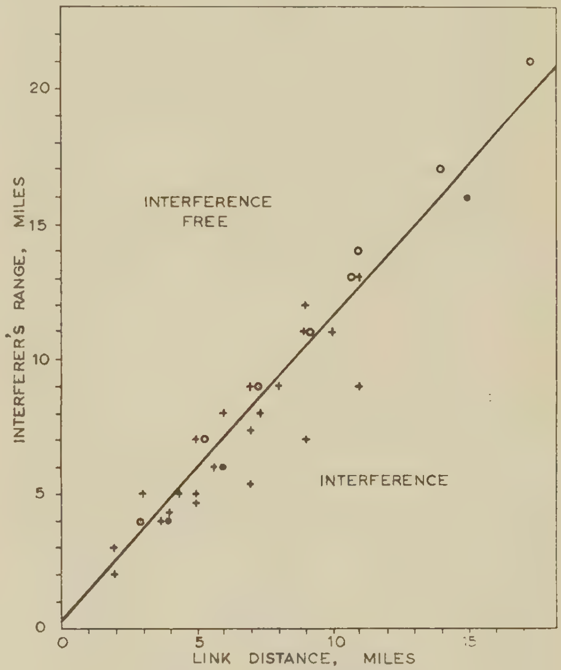


Fig. 5.—Voice-interference tests, showing how interferer's range varies with link distance over flat country at 59 Mc/s for the 15-watt set.

● No interference.
+ Interference, wanted link workable.
○ Interference, wanted link unworkable.

curve is a plot of interferer's range against link distance taking $I_c = 3$ dB and field strength for the various distances direct from Fig. 2. This provides an interesting comparison with the initial approach to the problem portrayed in Fig. 2 and the derived results which follow. For clarity, neither cases of heavy interference on the interference side of the curve, nor cases of non-interference on the interference-free side of the curve, have been shown.

(3.3) Field-Strength Distribution at a Given Distance

The analysis of field-strength distribution was performed on the observed data before conversion of the results obtained with the measuring set to microvolts per metre. This is valid, provided that the means of measurement are consistent, since the spread of results at each distance is being investigated here and not the variation of mean or median values with distance. Tests of the measuring equipment were carried out at intervals throughout the trials in the form of a re-calibration, and both

instruments were found to be consistent within approximately ±5%. Observations with the measuring set were treated as separate samples from those obtained using the field-intensity meter.

Preliminary investigation of the distribution of the field strength at each distance revealed a skew distribution with mean values in excess of median values.[19] The data were then fitted to a log-normal distribution using probability paper. It was found that the field-strength distribution approximated to log-normal at all ranges, the fit improving with sample size, and Fig. 6 clearly shows the excellence of fit. The examples shown

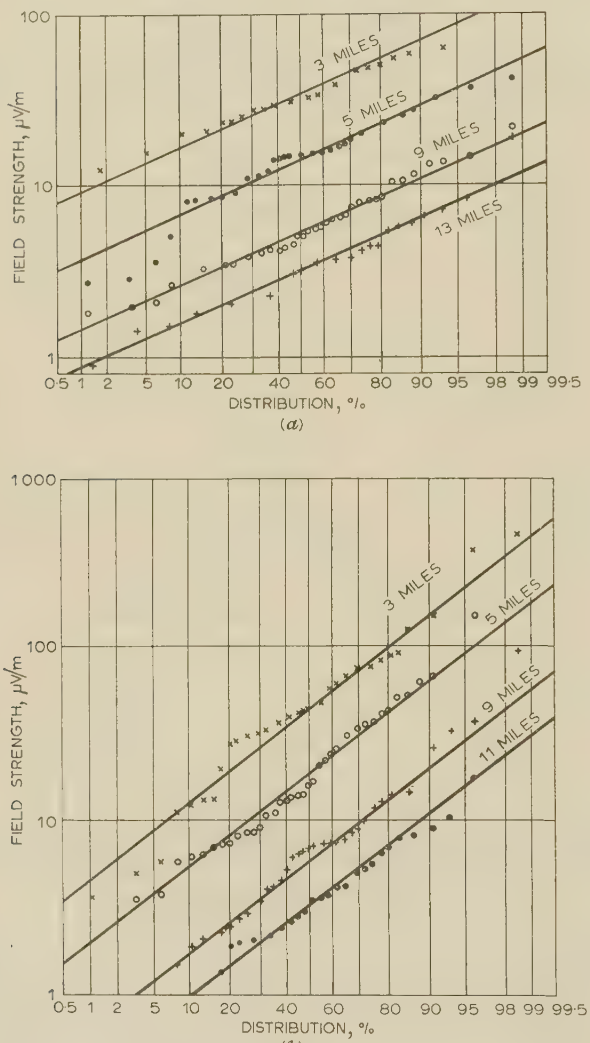


Fig. 6.—Distribution of field strength at given distances for 59 Mc/s; 15-watt set with 8 ft-rod aerial.

* Measuring set readings uncorrected.
(a) Flat country; curves indicate log-normal distributions; standard deviation, $s_F = 0\cdot28$.
(b) Hilly country; curves indicate log-normal distributions; standard deviation $s_H = 0\cdot465$.

are quite representative, and even truncated samples at the longer ranges, where it was often possible to detect that a transmission was being made but not possible to measure it, exhibit the same degree of fit. On this evidence it was accepted that field strength over a number of paths of a given distance over a particular type of terrain followed a log-normal distribution. All observed data were therefore converted to logarithms to the base 10, and subsequent analyses were carried out on the transformed variate.

This facilitated working in decibels and also made the ε of distribution of field strength independent of scale. Th strength distributions at the various distances were ob correlated and the independence of scale permitted comparison.

The samples at each distance for each method of measu and type of terrain were analysed separately, and th parameters for samples at 59 Mc/s, where 90% or m transmissions were measured, are given at Table 2. By tion alone it would appear fair to assume that the varia log field-strength tend to a particular value for each country independent of range, method of measurem transmitting aerial. The significance of this hypothe tested by Bartlett's test for variability.[20, 21] When all s are considered the results indicate that the variability field-strength has a single value for each type of cou approximately the 20% level of significance for flat coun better than the 20% level of significance for hilly countr the samples with rod transmitting aerials measured w measuring set, the level of significance of a homogeneous v is considerably better than 50% for both flat and hilly c Some further analysis was carried out on transmissions re simultaneously on the field-intensity meter and measur This indicated that the distribution of field strength at distance is largely a function of terrain and that se variations and measuring inaccuracies are of sec importance.

The significant factors arising from this analysis follow:

(a) Over different paths of the same length in a giv of country the received field strengths, from a given aerial arrangement up to heights $1\cdot5\,\lambda$ above ground, a mate to a log-normal distribution, certainly up to r few miles in excess of the range of the radio horizon.

(b) This variation in field strength is largely depen the conformation of, and accretions to, terrain. order contributory factors are the condition of the atmospheric and cosmic conditions,[11] set perfo measuring accuracy and extraneous interference.

(c) The variability of the log-normal distributions strength at each distance appears to be, within practica the same for a particular type of country irrespective o up to ranges at a few miles beyond the radio horizon.

(d) For the two different types of country consid variances of the log-normal distributions of field stren significantly different. The standard deviations deriv the field strength survey are $s_F = 0\cdot28$, or 5·6 dB, country and $s_H = 0\cdot465$, or 9·3 dB, for hilly country.

The above results are for vertically polarized waves; f zontal polarization at these frequencies the results sh much the same, although the possibility of a greater va due to the difference in propagation of horizontally and v polarized waves over ridges should not be overlooked.[11,]

The results obtained at 38 Mc/s over flat country, a comparatively few, are not significantly different from 59 Mc/s, as shown in Table 3. Over flat country us aerials, with samples of 7 or 8 observations at ranges be and 15 miles, the evidence of common variability, s_F is within the 5% confidence limits.

It is noted that at both 38 and 59 Mc/s the variab ranges of 1 and 2 miles, with one exception, conformin general assumption of common variability for each country, tends to be greater than at longer ranges. Th has not been pursued, but it may well be due to the er

Table 2

FIELD-STRENGTH DISTRIBUTION DATA AT 59 Mc/s

:ountry	Distance	Flat country						Hilly country					
		8 ft-rod aerial			Elevated aerial			8 ft-rod aerial			Elevated aerial		
:trument		Sample size	Mean/Median	s	Sample size	Mean/Median	s	Sample size	Mean/Median	s	Sample size	Mean/Median	s
	miles		dB	dB		dB	dB		dB	dB		dB	dB
l-intensity eter	1	34	2·1	7·0	14	1·9	7·0	34	5·3	10·4	12	3·9	7·5
	2	31	2·6	7·2	13	2·3	7·0	32	3·3	9·6	10	2·6	6·9
	3	27	1·7	5·0	14	1·3	5·5	36	5·8	10·1	11	8·5	11·4
	4	31	1·3	5·1	15	1·7	5·9	29	5·1	9·1	11	4·1	8·4
suring set	3	43	1·7	5·7	17	1·2	5·1	42	4·2	8·9	16	3·0	7·9
	4	25	2·9	6·0	13	0·7	3·6	29	4·0	9·0	12	4·0	8·7
	5	40	1·5	5·4	13	2·3	7·4	44	4·3	8·6	19	3·5	8·7
	6	—	—	—	—	—	—	27	4·1	9·7	—	—	—
	7	41	2·4	5·8	15	2·3	6·9	47	4·5	9·3	16	7·9	11·7
	8	—	—	—	—	—	—	29	4·0	7·1	—	—	—
	9	41	1·4	4·9	17	2·6	7·8	41	5·3	9·4	15	5·2	11·0
	10	—	—	—	—	—	—	28	4·8	9·3	—	—	—
	11	41	2·4	5·7	15	0·7	4·0	39	5·1	8·4	16	6·8	10·6
	12	—	—	—	—	—	—	25	6·6	11·4	—	—	—
	13	37	1·7	5·2	14	3·4	7·4	41	5·4	10·0	16	3·8	8·6
	14	—	—	—	—	—	—	26	5·9	10·8	—	—	—
	15	38	1·5	5·5	24	2·4	7·1	—	—	—	16	7·7	12·0
	16	—	—	—	—	—	—	27	5·1	9·6	—	—	—
	17	35	0	6·4	22	1·4	5·4	34	5·3	9·8	20	6·3	11·6

s = Standard deviation about median field.

ow effect of obstacles close to the base site at the shorter es.

(3.4) Measuring-Set Aerial Factor

e factor H_1, associated with the measuring-set aerial, was mined and found to vary between 3 and 4·3 dB, depending ie method of derivation and type of country, the lower s being associated with hilly country. A comparison of ltaneous measurements of the same transmissions by both intensity meter and measuring set gives $H_1 = 4{\cdot}3$ dB with orrelation coefficient of 0·558 for flat country and = 3·49 dB with a correlation coefficient of 0·402 for hilly try. The degree of association between the paired readings t as high as might be expected for two reasonably consistent iments measuring the same transmission. This is perhaps o surprising, since some of the samples were small and vo measuring aerials obviously could not be in exactly the place.* From a comparison of all readings at distances een 1 and 11 miles H_1 was found to be 4·2 dB for flat try and 3·05 dB for hilly country. The mean values of = 3·25 dB for hilly and $H_1 = 4{\cdot}25$ dB for flat country were to convert measuring set readings to microvolts per metre.

(3.5) Variation of Field Strength with Distance

om the detailed results of the survey, median field strengths lotted against distance in Fig. 7. Plotted on the same axes alculated field-strength curves derived from the approximate strength, E_0, over a spherical smooth earth[23] adjusted to E_f, edian field strength over irregular terrain at f megacycles cond, using the empirical correction presented by Saxton,[15]

$$E_0 \simeq \frac{7\sqrt{(PG)}}{d}\frac{4\pi h_T}{d\lambda}h_R \text{ volts per metre} \quad . \quad . \quad (1)$$

the paired reading samples had been larger it might have been possible to ate what might be termed 'within site' variation of field strength; this is, of included in the distribution of field strength at a given distance.[14]

Table 3

FIELD STRENGTH DISTRIBUTION DATA AT 38 Mc/s

Instrument	Distance	Sample size	s
	miles		dB
Field-intensity meter	1	7	13·5
	2	7	7·9
	3	7	4·0
	4	8	5·4
	5	7	5·7
Measuring set	2	7	4·4
	3	7	2·8
	4	8	3·7
	5	8	4·0
	7	8	7·7
	9	8	4·2
	11	8	3·1
	13	8	3·6
	15	8	4·8

Flat country, 8 ft-rod transmitting aerial.

$$20 \log_{10}\frac{E_f}{E_0} = [37 - 20 \log_{10} f] \text{ decibels} \quad . \quad . \quad (2)$$

It is clear from Fig. 7 that for all practical purposes the slopes of the calculated and experimental curves are identical within the normal working range of a set having a minimum usable signal of approximately 1 μV/m. In the calculation the gain of the 8 ft-rod vehicle-mounted aerial 3·9 m above ground has been taken as half that of a $\lambda/2$ dipole. The experimental curve over hilly country gives greater median field strength than the calculated curve. It is also apparent that the average advantage of using an elevated aerial in hilly country is less than in flat country, the results overall being indistinguishable from those

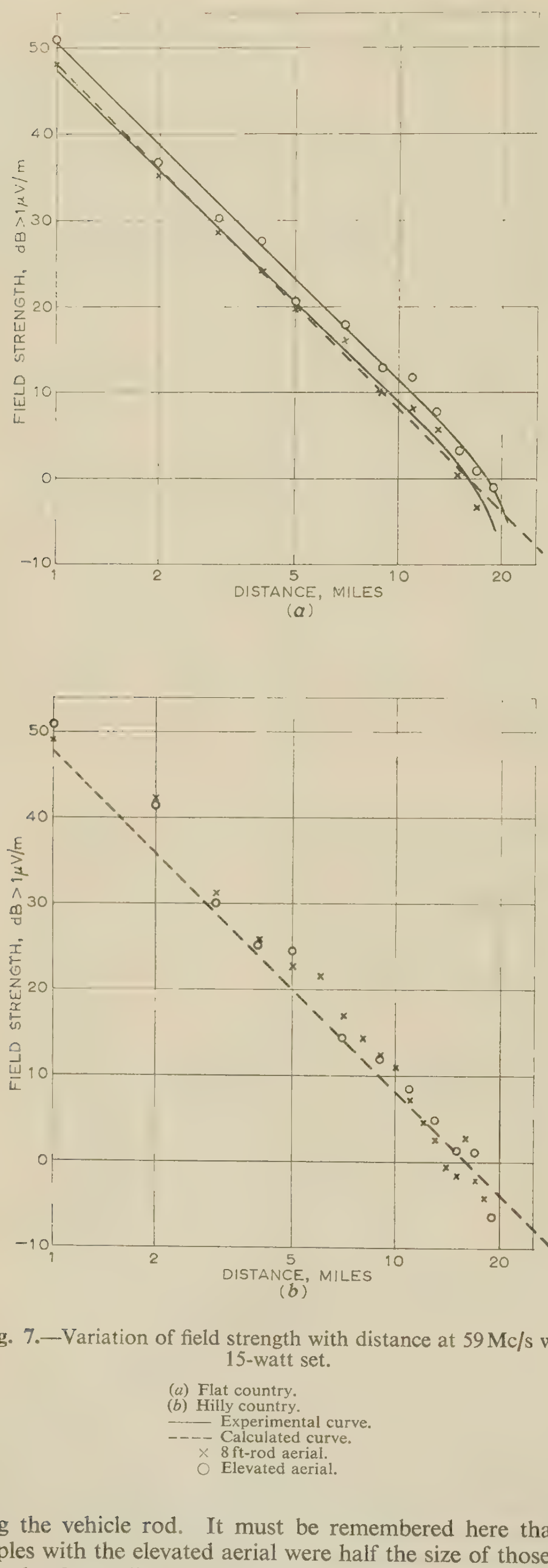


Fig. 7.—Variation of field strength with distance at 59 Mc/s with 15-watt set.

(a) Flat country.
(b) Hilly country.
——— Experimental curve.
- - - - Calculated curve.
× 8 ft-rod aerial.
○ Elevated aerial.

using the vehicle rod. It must be remembered here that the samples with the elevated aerial were half the size of those with the rod. Generally, calculated curves could be used to estimate median field strengths up to ranges where 1 μV/m is reached without appreciable error. Beyond this range the experimental[9, 11, 22, 23] curves of field strength fall off more rapidly than proportional to $1/d^2$.

(4) PROBABILITY OF INTERFERENCE

(4.1) General Case

It has been shown that the distribution of field stren[...] any given distance over a type of country approximates t[...] normal with a common variance for the type of country[...] spective of range. Consider the probability of interferenc[...] situation such as depicted in Fig. 1. At B, e_l is normally[...] buted about $\bar{e}_l$, with standard deviation s_l, and e_x about $\bar{e}_x$[...] standard deviation s_x. It can be shown that, if the distrib[...] of e_l and e_x are normal, the distribution of $(e_l - e_x)$ [...] normal about a mean value $\bar{u} = (\bar{e}_l - \bar{e}_x)$ with a variance[...]

$$s_u^2 = s_l^2 + s_x^2$$

But it has been shown that for a given type of count[...] variance may be considered to have a single value; theref[...] flat country,

$$\left.\begin{aligned} s_l &= s_F = s_x \\ \text{and} \qquad s_u &= s_F\sqrt{2} \end{aligned}\right\} \quad . \quad . \quad . \quad .$$

The probability of interference is given by the chance th[...]

$$u = e_l - e_x < i \quad . \quad . \quad . \quad .$$

Since u is normally distributed about $\bar{u}$ with a known sta[...] deviation s_u, for a given type of country, to find the prob[...] of $u < i$, convert $(i - \bar{u})$ into units of standard deviation[...]

$$X = \frac{i - \bar{u}}{s_u} = \frac{i - \bar{u}}{s_F\sqrt{2}} \quad . \quad . \quad . \quad .$$

for flat country

Similarly,

$$X = \frac{i - \bar{u}}{s_H\sqrt{2}} \quad . \quad . \quad . \quad . \quad .$$

for hilly country

The Tables for the normal curve[25] give the probability of[...] for any particular value of X, i.e. the probability of interfe[...] For example, if $l = 5$ miles and $x = 7$ miles over flat co[...] using the 15-watt set with rod aerials at 59 Mc/s, what [...] probability of interference?

From Table 1, $I_c \simeq 3$ dB, i.e. $i = 0{\cdot}15$ and $s_F =$ [...] therefore $s_F\sqrt{2} = 0{\cdot}396$.

From Fig. 7, $\bar{e}_l = 1{\cdot}0$ and $\bar{e}_x = 0{\cdot}72$; therefore $\bar{e}_l - \bar{e}_x =$ $0{\cdot}28$.

$$X = \frac{i - \bar{u}}{s_F\sqrt{2}} = -0{\cdot}33 \quad . \quad . \quad . \quad .$$

From the Tables of the normal curve[25] it is found tha[...] $X = -0{\cdot}33$, $62{\cdot}93\%$ $u > i$. Therefore there is about[...] probability of interference, or conversely a 63% chan[...] non-interference.

It should be noted that the above method of determini[...] probability of interference assumes no correlation betw[...] and e_x. If there is correlation, e.g. if e_l increases beca[...] ambient conditions, e_x is also likely to increase and the ch[...] of interference are reduced. It is thought that any such [...] tions would be insignificant in the presence of the prepon[...] influence, namely roughness of terrain.

(4.2) A Graph for Practical Use

For practical use a simple graphical presentation is [...] preferred. Eqn. (7) may be written

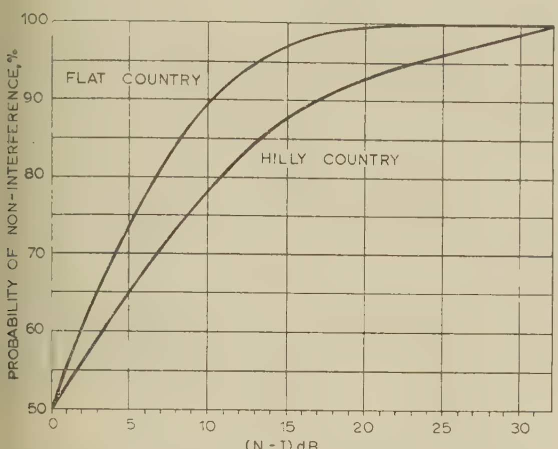
$$-s_u X = (\bar{u} - i) \quad . \quad . \quad . \quad .$$

in decibels this becomes

$$-20\,s_u X = 20\,(\bar{u} - i)$$
$$= 20\,(\bar{e}_l - \bar{e}_x - i)$$
$$= (N - I) \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

re N = median field strength from the wanted station minus ian field strength from the interferer.

or various probabilities X is given in the Tables and s_u has n found for flat and hilly country; hence graphs of $N - I$ nst various probabilities of non-interference can be plotted both types of country (Fig. 8).



8.—Probabilities of non-interference expressed as $(N - I)$ dB, i.e. the difference between wanted station field and the sum of interferer's field and acceptable protection ratio.

(5) PRACTICAL APPLICATIONS

(5.1) Estimation of the Probability of Non-Interference

he probability of non-interference between mobile stations now be easily assessed within the range of the experimental ι. Apart from curves of median field strength against ance and probabilities for non-interference against $N - I$, necessary to know the powers, P, aerial gains, G, and proion ratios, I, of the sets considered. Since received field ngth is proportional to $P^{1/2}G^{1/2}$, allowance for these factors be made directly on median field-strength curves.

he general procedure to estimate the probability of nonrference for a situation as in Fig. 1 is

(a) Determine the difference in decibels between wanted and interring station median field strengths from measured or calculated ırves.

(b) Subtract from this figure the acceptable protection ratio in ecibels.

(c) From Fig. 8 read the probability of non-interference correonding to this value.

procedure could, of course, be reversed, and starting with sired probability of non-interference, the distance acceptable veen stations for this probability of non-interference could be nated.

(5.2) Estimated Interference Areas

ithin the range where observed median field strength varies oximately as $1/d^2$, probabilities of non-interference estimated ı calculated and observed field-strength curves are for all tical purposes the same. Thereafter, the probabilities of interference based on the observed curves are greater, i.e. separation distances for a given probability of non-interference are less when estimated from the observed curves than when calculated curves are used. For the 15-watt set with 8 ft-rod aerial the common-channel separation is 20·5 miles for 90% probability of non-interference using the observed field-strength curve in Fig. 7(a). The corresponding distance using the calculated curve is 34 miles. At 34 miles only approximately 3% of the values of field strength from this set are likely to exceed 3 dB below the minimum usable signal (m.u.s.) even on the basis of the calculated curve, so that this separation distance is excessive. If the difference between the ranges to the interferer and the wanted station from the receiver for a given probability of non-interference is considered, it continually increases with range to the wanted station when calculated curves are used. However, if measured field-strength curves are used, this difference in ranges passes through a maximum when the range to the interferer approximates to a distance just beyond the radio horizon, as shown at Fig. 9. At this distance median field

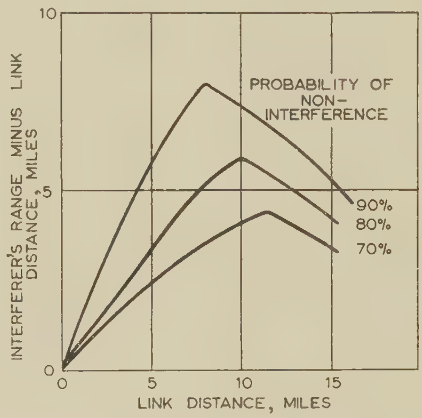


Fig. 9.—Range differences for given probabilities of non-interference.

from the interferer approximates to the m.u.s.[5] This maximum range difference may be used to estimate interference areas beyond the limits of the experimental data.

In normal practice interference areas should be estimated on the assumption that the wanted station may be received on the m.u.s. Therefore, in common-channel operation the frequency-separation distance is taken as the sum of the maximum difference between interfering and wanted station ranges and the range at which median field from the interferer reaches the m.u.s. For the 15-watt set this maximum range difference can be taken direct from Fig. 9. Alternatively, it can be estimated from calculated field-strength curves by assuming the interferer to be at the range to produce the m.u.s. at the receiver. From the appropriate value of $(N - I)$ decibels (Fig. 8) and the protection ratio the desired wanted station median field at the receiver can be determined and hence the range to the wanted station. For the 15-watt set with a rod aerial over flat country the maximum range difference for 90% probability of non-interference from the calculated curve in Fig. 7(a) is 16 miles (m.u.s. = OdB > 1 μV/m) minus 7·4 miles (N = 13·2 dB), i.e. 8·6 miles. The estimated frequency-separation distance for common-channel operation with at least 90% probability of non-interference is therefore 16 + 8·6 = 24·6 miles.

In practice the majority of stations would not be working on the m.u.s., and the repetition of frequencies outside the estimated common-channel interference area should result in virtually interference-free working. The methods outlined can be applied similarly to the determination of adjacent-channel-interference areas and other problems.

(6) CONCLUSIONS

The results of the field-strength survey at 59 Mc/s, at ranges up to 21 miles, show that the variation of field strength with distance is in close agreement with calculation. The analysis of the distribution of field strength at a given distance shows that it approximates closely to a log-normal distribution, which may be considered to have a single value of variance for a given type of country, irrespective of range. This distribution function was determined from trials where the transmitter siting was approximately random, and it includes position errors, set performance, measuring inaccuracies and ambient conditions, as well as the predominent effects of terrain and obstacles. Only vertical polarization has been considered, but the results for horizontal polarization should be similar. Based on data derived from the experimental results, probabilities of interference between mobile field radio sets in the v.h.f. band have been estimated.

(7) ACKNOWLEDGMENTS

Thanks are due to all who assisted in the manning and maintenance of the field equipment during the extensive field trials, often carried out in arduous conditions. In addition, acknowledgments and thanks are due to those who assisted in the considerable and often tedious task of routine computation.

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RADIO INTERFERENCE FROM IGNITION SYSTEMS

Comparison of American, German and British Measuring Equipment, Techniques and Limits

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(*The paper was first received* 15*th July*, 1960, *and in revised form* 11*th January*, 1961.)

SUMMARY

e paper records the results of some tests made in the United s to compare results obtained with American, German and sh radio-interference measuring sets when measuring the inter- ce radiated by the ignition systems of motor vehicles. The tests ' the close agreement between the results obtained when using the us national measuring sets and confirm the reliability and con- ncy of measuring equipment conforming to the specification of International Committee on Radio Interference (C.I.S.P.R.). ' also attempt to establish a relationship between the results of and quasi-peak measurements.

e paper also tabulates and discusses the test conditions specified e various countries and quotes the limits of ignition interference nmended or statutory in the countries concerned. A direct parison of these limits, which is made possible by these measure- s, suggests that the existing and proposed requirements of other ntries are more onerous than those of the United Kingdom.

(1) INTRODUCTION

he work described in the paper resulted from the discussions e 1958 meeting of the International Committee on Radio rference (C.I.S.P.R.) at which the United States delegation mitted a draft standard for radio interference from motor cles which had been prepared by the American Society of omotive Engineers (S.A.E.). This proposed standard set ts for the radiated interference which appeared to be more re than those in the United Kingdom and some of the other opean countries, and furthermore the American methods of surement and specification of limits were not in accordance C.I.S.P.R. practice.

order, therefore, to help the C.I.S.P.R. in its aim of eving unified methods of measurement and limits for radio interference, it was decided to make tests to compare American, German and British measuring equipment and techniques. These tests were made in the United States in 1959.

(2) MEASURING EQUIPMENT AND TECHNIQUES

In the United States, peak values of the interference are measured and the S.A.E. draft standard quotes limits in these terms, whereas equipment which conforms to the international C.I.S.P.R. specification measures quasi-peak values. The C.I.S.P.R. has made the recommendation that meters other than the quasi-peak type could with advantage be added to measuring sets, but has not so far specified the characteristics of these meters. The limits in the British and, in general, all European standards quote quasi-peak values, although in some cases peak values are also given.

The S.A.E. draft standard recommends two measuring sets as suitable for the measurement of ignition interference, type A which measures peak values only and type B which measures both peak and quasi-peak.

The German set measures both peak and quasi-peak, but the standard British set measures quasi-peak only. An earlier version of the German set has been described in an E.R.A. report.[1]

Two type A and one type B American sets, one German set and two British sets were used in the tests.

The main characteristics of the sets and the conditions of use are given in Table 1.

Table 1

CHARACTERISTICS OF MEASURING SETS

Set	Frequency range	Bandwidth	Output meter	Time-constants for quasi-peak: Charge	Time-constants for quasi-peak: Discharge	Aerial	Height of aerial above ground	Distance of aerial from vehicle	Polarization
	Mc/s	kc/s		ms	ms				
nerican type A ..	30–400	100	Peak	—	—	$\lambda/2$ dipole	7·5 ft	50 ft	Horizontal
nerican type B ..	30–400	100	Peak and quasi-peak	1	600	$\lambda/2$ dipole	7·5 ft	50 ft	Horizontal
rman	30–300	120 on quasi-peak 150 on peak	Peak and quasi-peak	1	500	$\lambda/2$ dipole	3 m (10 ft)	10 m (33 ft)	Corresponding to maximum value
itish	30–220	100	Quasi-peak	1	500	$\lambda/2$ dipole	$\not<$2 m	33 ft	Vertical

The S.A.E. draft standard calls for measurements to be made from the two sides, front and back of the vehicle but in practice this is reduced to one measurement from the nearside. The engine is run at a steady speed of about 1 000 r.p.m., which is high enough to cause operation of the generator-regulator unit. The aerial is mounted above a radial ground plane which consists of eight equally spaced copper wires each 16·5 ft long and jointed together at a point directly below the centre of the dipole aerial. In Germany the measurements are made at the front

itten contributions on papers published without being read at meetings are d for consideration with a view to publication.
: paper is based on Report Ref. M/T 138 of the British Electrical and Allied tries Research Association.
Ball is with Joseph Lucas (Electrical), Ltd.

of the vehicle and in Britain from the nearside. In neither country is a ground plane used. The British practice is to accelerate the engine slowly through its speed range, the maximum value of the interference being recorded.

(3) TEST PROCEDURE

The site of the measurements was the Delco-Remy testing station, which is a few miles outside Anderson, Indiana, in an interference-free area. A number of vehicles of different makes, both American and European, were assembled for the tests, which were attended by representatives from American car and ignition-equipment manufacturers, measuring-set manufacturers and the Federal Communications Commission, as well as three representatives from Germany and the authors from the United Kingdom.

Comparisons were made on peak and quasi-peak measurements between the various measuring sets on the horizontally- and vertically-polarized components of the fields and at distances of 33 and 50 ft from vehicle to aerial. Generally the engines were run at a steady speed of about 1 000 r.p.m., but some tests were made with the engines rapidly accelerated from low speed.

A radial ground plane was placed at a convenient point in front of the test hut and the various vehicles were positioned broadside-on at the appropriate distances. For the whole of the tests the aerial was positioned over the ground plane. The cars used during the tests ranged from small 4-cylinder to large 8-cylinder models.

The measuring sets were also compared on the field radiated by a pulse generator, the amplitude and repetition rate of which were accurately controlled and maintained constant.

(4) TEST RESULTS

For ease of comparison all the values given in this Section are in relation to a 100 kc/s bandwidth; in the case of the German set, where the bandwidth is not 100 kc/s, the appropriate bandwidth corrections have been made to the readings.

Table 2 shows the comparison of peak readings for th American type A and the type B sets as obtained on a 19 saloon car fitted with resistive ignition cable. Unfortu it was not possible to make simultaneous readings wit

Table 2

COMPARISON MEASUREMENTS WITH AMERICAN TYPE A TYPE B MEASURING SETS

Frequency	Levels in dB above 1 μV/m		
	Set 1	Set 2	Set 3
Mc/s	dB	dB	dB
30	19·5	20·5	19
40	14·5	13	15·5
50	22·5	22	19
60	32·5	32	29·5
70		25·5	
80		26·5	24·5
90		37·5	29·5
100	19	28·5	19
120	29·5	39·5	29
140	38	37·8	37
160	24	30	28·5
180		32·5	35·5
200	39	40	37·5
220	27	30	31
240	34	38·7	39·5
260	37·5	45·5	39
300	40	41	43·5
350	33·5	39·5	37
400	35	37·5	37·5

Measurements were made on a 1959 V8 4-door saloon fitted with resistive cable. Distance of aerial from vehicle, 50 ft. Horizontal dipole aerial, 7·5 ground.

Sets 1 and 2, American type A.
Set 3, American type B.

Table 3

RATIO OF PEAK TO QUASI-PEAK VALUES AND OF DISTANCE FROM VEHICLE TO AERIAL OF 33 FT TO 50 FT AS MEASURED ON GERMAN SET

Frequency	Levels in dB above 1 μV/m														
	33 ft distance						50 ft distance						Ratio 33 ft/50 ft		
	Peak		Quasi-peak		Ratio		Peak		Quasi-peak		Ratio		Peak		Quasi-
	V	H	V	H	V	H	V	H	V	H	V	H	V	H	V
Mc/s															
35	42	25	25	14·5	17	10·5	39	27	22	16	17	11	3	−2	3
50	40	29	22	4	18	25	38	25	14	−2	24	27	2	4	8
70	42	36	24	15	16	21	38	31	20	12	18	19	4	5	4
85	44	35	27	15	17	20	41	34	23	12	18	22	3	1	4
100	45	34	24	18	21	16	38	28	21	3	17	25	7	6	3
150	52	44	28	28	24	16	47	33	28	19	19	14	7	11	0
200	50	56	31	18	21	38	46	44	21	19	25	25	6	12	10
250	50	44	35	31	15	13	41	44	31	21	10	23	9	0	4
300	41	53	23	27	18	26	43	40	12	12	21	18	−2	13	11
Mean values ..					18·6	20·6					18·8	20·4	4·3	5·6	5·2

Overall mean value of peak to quasi-peak, 19·6 dB.
Overall mean value of ratio of levels at 33 ft to 50 ft, 5·4 dB.
Measurements made on 1959 V8 saloon, fitted with resistive ignition cable. Height of aerial above ground, 10 ft.
V Vertical polarization.
H Horizontal polarization.

: sets, and in fact quite long intervals sometimes occurred een the readings of one set and another. The engine was at a steady speed of about 1 000 r.p.m. at no load. This is an ideal condition of engine operation, and it was found ssary from time to time to give the car a brisk run on the in order to clean-up the engine. Although the readings of hree sets often agreed, differences of up to 10 dB sometimes rred and it is possible that some of them were due to irregu-es in the interference levels.

ible 3 shows the results of measurements with the German on the V8 car. Here peak and quasi-peak values were rded at distances of 33 and 50 ft for both vertical and zontal polarization. The peak/quasi-peak ratios show

Table 4

ATIO OF PEAK TO QUASI-PEAK VALUES AS MEASURED ON AMERICAN TYPE B SET

Frequency	Levels in dB above 1 μV/m		
	Peak	Quasi-peak	Ratio
Mc/s	dB	dB	dB
80	24·5	4·5	20
90	29·5	8·5	21
100	19	6·5	12·5
120	29	12	17
140	37	15·5	21·5
160	28·5	15·5	13·0
180	35·5	16	19·5
200	37·5	14·5	23
220	31	16	15
240	39·5	18·5	21

an value of peak/quasi-peak ratio, 18·3 dB.
asurements made on 1959 V8 saloon, fitted with resistive ignition cable. Hori-l dipole aerial 7·5 ft above ground and 50 ft distant from vehicle.

iderable scatter—again, in the authors' opinion, largely butable to variations in the interference levels, with a mean e of 19·6 dB. Similarly, there is a large scatter in the es for the ratio of the levels at 33 ft to those at 50 ft. The ad of results from an apparent gain of 2 dB to an attenuation 3 dB with a mean value of 5·4 dB attenuation is entirely in rdance with the experience of the E.R.A., which is that, although individual determinations are unreliable, a sufficiently large number of readings gives a mean value of just under 6 dB for the attenuation from 33 to 50 ft. Up to 150 Mc/s the levels for vertical polarization are significantly higher than those for horizontal polarization, and British experience as a result of thousands of tests is that this is generally true.

Table 4 shows the comparison between peak and quasi-peak levels from the V8 car as recorded by the American type B set for horizontal polarization and at the distance of 50 ft. Again there is a large scatter in the results, but the mean value of 18·3 dB for the peak/quasi-peak ratio is close to the value obtained with the German set.

Results for the two British sets on the V8 car are shown in Table 5. Measurements were also made of the horizontally-polarized field at 33 ft and of both polarizations at 50 ft, but the levels so obtained were comparable with, or even below, the set noise. They were therefore meaningless and so have been omitted. The agreement between the two sets is good; the average difference is no more than about 1 dB, and only in two cases is there a relatively large difference of 5 dB and this was probably due to variation in the interference level.

Comparison measurements on peak and quasi-peak values with the American type B, German and British sets were made

Table 5

COMPARISON MEASUREMENTS WITH BRITISH MEASURING SETS

Frequency	Levels in dB above 1 μV/m	
	Set 1	Set 2
Mc/s	dB	dB
50	23·5	22·5
80	25	26·5
100	21·5	26·5
120	23·5	24·5
140	26·5	31·5
160	29·5	32·1
200	30·8	29·0
220	29·3	28·3

Measurements made on 1959 V8 saloon, fitted with resistive ignition cable. Height of vertical dipole aerial above ground, 7·5 ft. Distance of aerial from vehicle, 33 ft.

Table 6

PEAK AND QUASI-PEAK MEASUREMENTS ON NEW 6-CYLINDER SALOON CAR FITTED WITH SINGLE 10 kΩ RESISTOR IN COIL LEAD

Frequency	Levels in dB above 1 μV/m							
	American type B		British		German		Ratio peak to quasi-peak	
			No. 1	No. 2				
	Peak	Quasi-peak	Quasi-peak	Quasi-peak	Peak	Quasi-peak	American type B	German
Mc/s	dB	dB	dB	dB	dB	dB	dB	dB
70	63·5	34·5	33·3	31·6	50	28·4	19	21·6
70*	54·4	27·1					27·3	
100	58·3	31·3	29·7	28	51	30·4	27	21·4
150	60	36·3	35	34	60	40·4	24	19·6
200	62·5	38·5	38	38	62	41·4	24	20·6

* Levels remeasured 1½ h after first readings.

Mean values of peak/quasi-peak ratios, 24·3 dB for American type B, 20·8 dB for German.
Horizontal dipole aerial 8·5 ft above ground and 33 ft from vehicle.

on a 6-cylinder saloon car fitted with a single 10 kΩ resistor in the coil-distributor lead. The results are given in Table 6, and here again it is seen that the agreement between the various sets on both peak and quasi-peak is generally good. The mean values of the peak/quasi-peak ratio of 24·3 and 20·8 dB for the American and German sets respectively are reasonably close to those previously given.

In an effort to remove the inaccuracies caused by the variation in the interference levels from the motor cars, comparison measurements with the American type B, German and British sets were made on the field radiated from a pulse generator, the output of which could be maintained at a constant level. Since one of the main objects of the tests was to compare the peak and quasi-peak methods of measurement, it was decided to make no further tests with the American type A sets but to us calibrating pulse generator from one of them as a constant s of noise. This generator was fed to a doublet aerial which positioned 20 ft from the aerial of the measuring set. results are given in Table 7 and it is seen immediately ho agreement between the sets improves with the consistency radiated field. The difference in peak readings betwee American type B and the German set in no case exceeds 0· which is extremely good when it is remembered that the Ame set has a slide-back peak meter whereas in the German s peak output is assessed visually by comparison with a brating pulse on the screen of a cathode-ray tube. The ment between the four sets on quasi-peak measurement is equally good since the overall variation between them i

Table 7

Comparison Measurements with Standard Pulse Noise Generator of American Type A Set Fed into an Aer

Frequency	Pulse repetition rate	Levels in dB above 1 μV/m					
		American Type B		German		British	
						No. 1	No. 2
		Peak	Quasi-peak	Peak	Quasi-peak	Quasi-peak	Quasi-peak
Mc/s	p.p.s.	dB	dB	dB	dB	dB	dB
100	50	52·8	42·7	52·4	44·1	41·3	42·2
100	100	52·8	45·3	52·4	44·6	44·6	44·7
100	200	52·8	45·8	52·4	45·8	45·8	45·7
100	1 000	52·8	46·3	52·4	46·8	46·8	46·2
150	50	41·0	33·7	41·5	34·5	34·5	35·7
150	100	41·0	36·0	41·5	35·0	35·0	36·7
150	200	41·0	36·0	41·5	35·5	35·5	37·2
150	1 000	41·0	36·5	41·5	35·7	35·7	37·7
200	50	44·5	37·5	45·0	37·8	37·8	39·7
200	100	44·5	40·0	45·0	38·8	38·8	40·7
200	200	44·5	40·3	45·0	38·8	38·8	40·7
200	1 000	44·5	40·5	45·0	39·3	39·3	41·2

Distance between radiating aerial and aerial of measuring set was 20 ft.

Table 8

Measurements on a Group of 1958 and 1959 Private Cars

Car	Frequency	Levels in dB above 1 μV/m							
		American Type B		German		British Quasi-peak		Ratio peak to quasi	
		Peak	Quasi-peak	Peak	Quasi-peak	V	H	American Type B	Ge
	Mc/s	dB	dB	dB	dB	dB	dB	dB	d
6-cylinder saloon (resistive cable)	70	44·4	24·4	41·3	21·7	14	16	20	19
	200	49·5	27·5	47·4	15·4	30	<25·5	22	3
V8 saloon (10 000 Ω resistor in coil lead)	70	69·4	47·4	66·0	48·1	47·6	48·6	22	17
	200	59·5	37·5	58·6	37·4	40	39	22	21
V8 saloon (resistive cable)	70	38·4	20·4	38·2	20·7	23·6	20	18	17
	200	44·5	25·5	42	17·9	30	26	19	24
4-cylinder saloon (10 000 Ω resistor in coil lead)	70	52·9	28·4	52·5	27·4	39	28	24·5	25
	200	63·5	40·5	62·2	40·9	38·8	40·8	23	21
4-cylinder saloon (no suppression)	70	48·4	27·4	41·6	24·8	43	30·6	21	16
	200	62	38·5	64·3	41·3	46·8	40·8	23·5	23
4-cylinder estate car (10 000 Ω resistor in coil lead)	70	48·4	23·4	43·2	24	32·1	24·1	25	19
	200	65·5	41·5	63·8	40·4	47·3	39·8	24	2

Mean value of peak/quasi-peak ratio, 22 dB for American type B set.
Mean value of peak/quasi-peak ratio, 21·8 dB for German set.
Dipole aerial 8·5 ft above ground and 33 ft from vehicle. Horizontal and vertical components of the field measured on the British set. Horizontal only on other

e than about 2 dB. In this connection it is interesting to › that, although the two British sets are about seven years their stability is such that very little divergence from the inal calibrations has taken place.

able 8 shows the results of measurements with the four sets a number of cars fitted with various types of suppression. h one or two exceptions, probably due to irregularities in the rference levels, the agreement between the sets is satis-orily close. There appeared to be no significant change in peak/quasi-peak ratio with the type of vehicle; the mean ies were 22 dB for the American and 21·8 dB for the German

he overall means of all the peak/quasi-peak ratios given in les 3, 4, 6 and 8 are 20·4 dB for the German and 21·6 dB for American set. These are also in close agreement. The ct of rapidly accelerating the engines of the cars from low ›d was not very marked; sometimes there was no change and 1o case did the level of interference increase by more than ut 2 dB over that produced by the steady running at 0 r.p.m.

(5) DISCUSSION OF RESULTS

he most important result of the Anderson tests was the eness and consistency of the agreement between the American, rman and British sets. Prior to these tests the opinion had n expressed in some quarters that the quasi-peak method neasurement could not be expected to be as consistent and able as the peak method, but the tests have completely dis-ved this. The sets in general agreed with each other to 1in 2 or 3 dB, and where differences appreciably greater urred they were in all probability due to changes in the inter-nce level between one measurement and the next.

s to the correlation between measurements at 33 and 50 ft, can be achieved without great error by assuming a 6 dB rease in level as the distance of aerial to vehicle is increased n 33 to 50 ft. Probably the exact figure is rather less than 3; a more accurate value would emerge if more measurements e made. Although no tests were made with the specific ntion of investigating the effect of aerial height above ground, impression, which is reinforced by previous experience, is t differences in the range 7–10 ft have insignificant effect on the rference levels recorded. This should therefore present no at difficulty when seeking international agreement on methods neasurement.

he problem of polarization, however, is not so simple. ce the television transmissions in the United Kingdom are inly vertically polarized, the vertical component of the inter-nce field is measured, whereas in the United States, with its izontally-polarized transmissions, only the horizontal com-ient is measured. Unfortunately there is no simple correla-1 between the horizontal and vertical components of the ition radiation as the relation between them is a complex and redictable function of frequency. In the authors' experience vertical component is usually the greater, at least up to id III frequencies.

rom the international aspect it may therefore be necessary to isure both the vertical and the horizontal components of the 1 as is the current practice in Germany.

he value of the peak/quasi-peak ratio of about 21 dB is not easonable on theoretical grounds and, moreover, is in close eement with that obtained previously by the E.R.A. from tests vehicles with the British set and an earlier version of the erican type B set. However, the opinion has been expressed correlation between peak and quasi-peak measurements in tion to engine speed is not possible since the two meters ond differently as the speed changes.

For instance, measurements in the United States are made at engine speeds in the region of 1 000 r.p.m. where the peak reading is at a maximum, whilst in the United Kingdom the engine is slowly taken through its speed range. It is argued that, as the quasi-peak reading is proportional to the product of peak value and repetition rate, the British method of recording the interference will produce a level which is constantly increasing with engine speed and hence correlation is impossible. The pro-portionality of the product of peak value and repetition rate holds for defined pulses such as are given by pulse-generator calibrators, but the ignition pulse is much more complicated since multiple sparking, variation of sparking-plug voltage between cylinders in multi-cylinder engines and change in sparking voltage with engine speed have to be taken into account. Because of the ignition-advance characteristics there is a compensating factor in that as the engine speed rises the sparking voltage falls and so the quasi-peak reading does not increase proportionally with engine speed. In fact, the results of over 1 000 measurements made by the British Post Office[2] on eight vehicles, each sup-pressed in several ways, show very little change in the quasi-peak value as the engine speeds change from about 1 000 r.p.m. to maximum speed. The average increase is not more than 3 dB and is appreciably smaller from 1 500 r.p.m. to maximum speed.

These results, together with those of the Anderson tests, suggest that for practical purposes the ratio of peak to quasi-peak measurements at a steady engine speed of about 1 000 r.p.m. can be taken as 20–21 dB.

The S.A.E. draft standard requires the use of a ground plane, and, as mentioned earlier, all the tests at Anderson were made with one, but it is not used in Germany and the United Kingdom. It would be interesting to have information on its possible effect since tests made in the United Kingdom have so far failed to detect any change with a ground plane of the type specified in the S.A.E. draft standard.*

(6) LIMITS

The problem of obtaining internationally agreed limits is difficult since the conditions of broadcasting and telecommuni-cation can vary so much from one country to another. This is a matter for the C.I.S.P.R. Working Group on Ignition Inter-ference, and any detailed discussion on it is outside the scope of this paper, which is primarily concerned with the comparison and correlation of the methods of measurement in the United States, Germany and the United Kingdom. Nevertheless, it is instructive to compare the various national regulations and standards which are already in force or are to be introduced shortly, and Fig. 1 shows the limits referred to a quasi-peak measurement on a 100 kc/s bandwidth set with the aerial 33 ft from the vehicle (peak/quasi-peak ratio 20 dB, 33/50 ft ratio 6 dB), for the United States, Germany, the United Kingdom, Belgium, the Netherlands and France. The French regulations require the receiving aerial to be 10 m above ground and so it is difficult to make an exact comparison of them with those of the other countries but they are included for the sake of completeness.

It is apparent that the British requirements, both as regards actual levels and the frequency ranges over which they apply, are less onerous than most of the others, even allowing for the difference between the vertical and horizontal components of the field. The authors' opinion is that, even with elaborate forms of resistive suppression, it will be very difficult to secure 100% compliance with most of the foreign requirements. In the United Kingdom the adoption of suppression measures more

* Since this paper was written further measurements have been made, both here and abroad, of the peak/quasi-peak ratio and the effect of the ground plane. In all cases the effect of the ground plane was found to be negligible. Measurements by the authors on about 40 vehicles at an engine speed of 1 000 r.p.m. gave a mean value of 19·3 dB for the peak/quasi-peak ratio with a range of 15–24 dB.

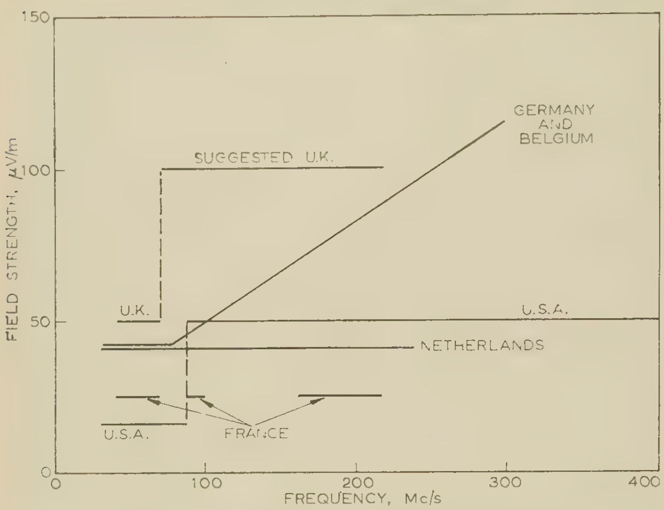

Fig. 1.—Comparison of field-strength limits.

Limits are referred to quasi-peak values, 100 kc/s bandwidth and 33 ft distance between vehicle and aerial.

elaborate than the use of $1 + n$ resistors, where n is the number of cylinders, or the equivalent in resistive cable would be regarded as unnecessary, both technically and economically, and with this form of suppression it is unlikely that a limit of 100 μV/m would be reached in all cases.

Another question is, how should the limits be applied to motor vehicles which are in large-scale production? In the United Kingdom a statistical approach is considered essential and for compliance with the regulations are based on a samp procedure, but this problem has so far received little attentic other countries.

(7) CONCLUSIONS

Measurements of ignition interference can be made equal accuracy in the United States, Germany and the U Kingdom, and the reliability and consistency of measu equipment conforming to the C.I.S.P.R. international spec tion have been confirmed.

In the opinion of the authors the possibility of correlating and quasi-peak measurements has been established.

The existing and proposed requirements of other cou for the control of ignition interference are more onerous those in the United Kingdom, and 100% compliance with by resistive suppression will be difficult and in some cases even be impossible.

(8) ACKNOWLEDGMENTS

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B.B.C. SOUND BROADCASTING 1939–60

A Review of Progress

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(1) INTRODUCTION

1938, Sir Noel Ashbridge, in presenting a review of broad-ng and television to The Institution,[1] was able to summarize teady and uninterrupted progress of broadcasting since the days. The period that followed was a momentous one in d history and brought with it important changes in broad-ng throughout the world.

ie present review deals mainly with progress made by the C. in sound broadcasting during the years 1939–60 inclusive, ugh reference has been made where appropriate to other r developments in this field both at home and abroad.

ie development of television will be reviewed in a companion r.

(2) DEVELOPMENT OF B.B.C. SERVICES

ie period under review started when the fear of another d war was uppermost in people's minds and when the urces of broadcasting had already begun to be used in many for influencing the minds and thoughts of listeners in r countries. The inevitable result was that those countries, iding the United Kingdom, were forced to combat this paganda by the extended use of broadcasting to present the 1 as objectively and as effectively as possible. The B.B.C. ed in January, 1938, the first of its foreign-language services essed to listeners abroad, in Arabic. The scope and volume iese services were greatly expanded soon after the outbreak ar; there was also a big expansion in the short-wave services essed to English-speaking audiences. The pattern of home idcasting was also radically changed during war time. The time engineering activities of the B.B.C. have already been subject of a paper presented to The Institution[2]; the develop-t of the broadcasting service in this country was also the ect of the Inaugural Addresses by Sir Noel Ashbridge,[3] ident of The Institution in 1941 and by Sir Harold Bishop,[4] ident of The Institution in 1953 (see also References 5 6).

1 1939, prior to the war, the B.B.C. offered to listeners at ie two alternative programmes, then known as the National Regional programmes, for a total of 202 hours per week. the outbreak of war, a single programme was introduced and transmitting stations were divided into two groups, one ring the South of England and South Wales and the other North of England and Wales, Scotland and Northern Ireland. transmitters in each group were synchronized on one wave-th as part of a plan to prevent their use by enemy aircraft avigational purposes, and high-accuracy crystal drive equip-t was developed in order that close synchronization could naintained without the use of line links. As the war pro-sed the service was improved by the building of additional ips of synchronized stations, one of which consisted of ransmitters mostly of 100 or 250 W, all operating on the wavelength and installed in large towns throughout the country. The number of transmitters reached its peak at the end of 1943, by which time two programmes, the Home Service and the Forces Programme, were in operation. There was then a total of 25 transmitters with an aggregate power of 800 kW, in addition to the 60 low-power transmitters mentioned above.

There were also six medium-wave transmitters broadcasting the B.B.C. European Service with a total power of 1 280 kW, including a new transmitter with a power of 400 kW which had been installed at Droitwich. In 1943, also, the B.B.C. completed another station capable of an output to the aerial of 800 kW at Ottringham, near Hull. At this station, four 200 kW transmitters were installed together with combining circuits, enabling them to be operated in parallel. It was decided to use three of them in parallel for the European Service, giving an output of 600 kW on a wavelength of 1 200 m; the fourth transmitter was used in the main North of England Home Service group.

Even greater expansion took place in the short-wave stations. At the outbreak of war, the B.B.C. had a single short-wave transmitting station at Daventry. This contained the two original low-power transmitters with which the Empire Service was started in 1932, the experimental low-power transmitter G5SW which had been transferred from Chelmsford, and three high-power transmitters. During the war a further five high-power transmitters were installed at Daventry and three further short-wave stations were built at Rampisham in Dorset, Skelton in Cumberland, and Woofferton in Shropshire. These contained an additional 22 high-power transmitters, six of them equipped with two entirely separate r.f. channels, each of which could be set up on a separate wavelength and operated independently. A further three high-power short-wave transmitters were provided at existing medium-wave stations, one of them by modifying a medium-wave transmitter not required during the war for its normal purpose.

A very big programme of aerial design and construction, including supporting masts and feeders was undertaken, and outdoor feeder switching towers were designed for the rapid connection of a transmitter to one of a number of aerials.

Some 150 war-time studios spread over the country were brought into service. There was a big increase in the use of recording because of the difficulty in assembling artists at certain times, particularly during air raids, and because of the need to repeat programmes in the Overseas Services. Entirely new disc recording and reproducing equipment was designed, including a portable disc recorder for the use of correspondents in the field. At the request of the Government, a monitoring service was set up which listened to broadcasts from all parts of the world.

When the post-war pattern of broadcasting was introduced in July, 1945, there were again two alternative programmes, the Home Service (with its regional variants) and the Light Programme. The Third Programme was added in September, 1946, and Network Three, using the same technical facilities at a different time of the day, in September, 1957. At the end of 1959 there was a total of 288 hours of sound broadcasting each week for listeners at home.

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The coverage of these programmes on the long and medium wavelengths in percentages of the total population is as follows:

Home Service: 93%.
Light Programme: 99%.
Third Programme and Network Three: 69%.

These figures refer to night-time reception and indicate the percentage of the population of the United Kingdom that can expect to obtain reception substantially free from fading. They do not take into account foreign interference, which seriously reduces the coverage at night, particularly of the Home Service. To combat this interference and extend the coverage as far as was practicable, 12 additional low-power medium-wave transmitting stations were built and brought into service during the years 1951–54.

It had long been foreseen by the B.B.C. that increasing congestion in the medium- and long-wave bands would cause reception conditions to deteriorate and would limit the total coverage obtainable. Towards the end of the war, therefore, the B.B.C. decided to carry out field trials of a system of broadcasting on ultra-short wavelengths using wide-band frequency modulation. The use of such a system had, in fact, been envisaged before the war. Frequency modulation was not new—in fact, a patent had been taken out in 1902—but it was not until Major Edwin Armstrong in America pointed out in 1936 the importance of wide-band frequency modulation on ultra-short wavelengths, using a noise-suppressing limiter in the receiver, that the possibilities of this system for broadcasting were realized.

The B.B.C. field trials began in June, 1945. Two 1 kW f.m. transmitters operating in the 45 Mc/s band were built in the B.B.C. Research Department, and one was installed at Alexandra Palace and the other near Oxford. The latter transmitter was later moved to Moorside Edge (near Huddersfield) to gain experience of reception in hilly country. The Alexandra Palace transmitter radiated the Home Service programme during the evening on 46·3 Mc/s and the same programme was radiated simultaneously by the Alexandra Palace television sound transmitter on 41·5 Mc/s, using amplitude modulation, for comparison purposes. Subsequently, tests were carried out from Alexandra Palace on 90·3 Mc/s, again using frequency modulation, as it seemed likely that a higher frequency band would be allotted to f.m. broadcasting if it were introduced in this country.

The results of these tests were so promising[7] that the B.B.C. decided to build a high-power transmitting station so that a full-scale field survey could be carried out which would indicate the coverage and grade of service to be expected and resolve the controversial 'a.m. versus f.m.' question.

In a broadcasting system, it is particularly important to keep the cost of the listener's receiver as low as possible. Bearing this in mind, only three systems merit consideration:

Conventional amplitude modulation (a.m.) as used in medium- and long-wave broadcasting.

Amplitude modulation with a noise limiter incorporated in a wide-band receiver (a.m.l.) (for the majority of the tests the bandwidth of the receiver was the same as that of the f.m. receiver, i.e. ± 75 kc/s).

Frequency modulation (f.m.).

The experimental transmitting station was built at Wrotham, Kent, and the tests began in July, 1950. A comprehensive series of field-strength measurements was made on these transmissions in order to check theoretical computations, following which a listening survey was carried out with the assitance of listeners provided mainly with specially designed v.h.f. receivers capable of receiving a.m., a.m.l., and f.m. transmissions.

The results of these tests clearly demonstrated the advantages of using frequency modulation.[8] The B.B.C. therefore prepared a plan for a v.h.f. f.m. broadcasting service to cover the part of the United Kingdom on the basis of three progra (Home, Light and Third) for each service area. For c areas this entailed some considerable overlapping becau the need to provide regional programmes appropriate t areas concerned.

The B.B.C. plan was submitted to the Postmaster Gener approval and referred by him to the Television Advisory mittee. The Committee in its report[9] on this subject r mended the adoption of frequency modulation for a sound-broadcasting service in this country. The Postm General subsequently gave his approval and the B.B.C. be implement its plan. The Wrotham station was brought in programme service in May, 1955, using v.h.f. f.m. transm

The primary object of introducing the v.h.f. service combat foreign interference and to extend the coverage B.B.C.'s three domestic programmes to areas where re from the medium-wave and long-wave stations was u factory. Other important advantages of this system a comparative freedom from most kinds of interference a improved quality of reproduction obtainable. At prese B.B.C. has in operation 20 v.h.f. transmitting stations gi coverage of more than 97% of the population.[10] F stations under construction or planned will increase the co to more than 98% (see Appendix 24.3).

The majority of these stations radiate the three hom grammes; the first four-programme station was ope Sandale in Cumberland in August, 1958, radiating the of England and Scottish Home Services as well as the Lig Third Programmes. Another 4-programme station of power is in operation at Wenvoe, near Cardiff.

Following earlier experiments, the B.B.C. began in Oc 1958, a series of experimental stereophonic transmissions. are broadcast on alternate Saturday mornings, using th work Three transmitters (v.h.f. and medium-wave) for th hand channel and the television sound transmitters fo right-hand channel. These transmissions were arrang meet the widespread public interest in the better-quality duction of sound and in the potentialities of stereo broadcasting. The B.B.C. is actively investigating p systems which would enable stereophonic programmes broadcast satisfactorily from a single v.h.f. transmitter w unduly impairing ordinary monophonic reception.[11] A vention on Stereophonic Sound Recording, Reproductio Broadcasting was held by The Institution in March, 1959.

(3) PROBLEM OF FREQUENCIES

The last 20 years have seen an enormous increase in t of radio frequencies for broadcasting purposes. In 19 medium-wave broadcasting bands were already becomin crowded. Since then the use of these frequencies for casting has still further increased, both in the number and of transmitters used. H.F. broadcasting has develope mously, while the v.h.f. band that was just coming into television in 1939 is now extensively used all over the wor addition, v.h.f. sound broadcasting—virtually non-exis 1939—now occupies an important place in the frequenc cation tables.[13] Frequencies in the upper ranges of th band, for which virtually no valves or equipment were av in 1939, are now in widespread use for broadcasting; th band is now also used for television transmission in the States (and in Germany and Italy), and its use in the Kingdom is under consideration. To meet the requirem broadcasting some increase was made in the space al for it in the Atlantic City Plan of 1947 and some slight increase at the Geneva Conference in 1959.

nultaneously with this great increase in the demands for encies, there has been a considerable widening of know- of the properties of some of the frequency bands and their bility for various types of service. The propagation erties of some frequency bands have been found by ience to be substantially different from what had been ght in 1939. This has affected broadcasting both directly indirectly. The service range of high-power television ons has proved greater than was expected, and the effect regularities in the terrain has affected the higher frequencies e v.h.f. band less than was thought likely. On the other , interference from distant v.h.f. stations is more severe had been expected and the development of ionospheric er services in the v.h.f. band and tropospheric scatter ces in the u.h.f. bands has introduced a new source of ference with broadcasting services. In f.m. reception, the ved signals caused by multi-path propagation result in anted amplitude and phase modulation of the primary l and consequent distortion of the programme output of receiver. Experimental work carried out by the B.B.C. ved that a receiver which provides inadequate a.m. sup- sion is much more prone to this type of distortion and that listortion could be greatly reduced by suitable circuit design out adding appreciably to the complexity and cost of the ver. The results of this work were communicated to the industry, and details of an f.m. receiver of B.B.C. design published.[14]

ork carried out all over the world has improved the know- e of the propagation properties of the various bands of uencies over land and water, and of ionospheric reflection tropospheric refraction; these studies have led to an increase he accuracy of prediction of circuit conditions, and the .I.R. has established acceptable curves for most of these litions.

lethods of conserving the frequency spectrum,[15] including application of information theory to the problem, have ived considerable attention. Means have not been found applying them so much to broadcasting as to other services, ely because broadcasting differs from other radio services at the transmitting and receiving installations are not under mon control. While single-sideband transmission is general ther communication services, it has not yet been possible to y it generally to sound broadcasting services because of complexity of the receiver. Vestigial sideband transmission however, been partially applied to television signals with result that, whereas 405-line television in this country ired a 7 Mc/s channel before the war, it is now satisfactorily mmodated in a 5 Mc/s channel.

nprovements to valves for receivers and for transmitting pment have enabled greater use to be made of the higher uencies, thus tending to relieve the load on the more heavily gested lower frequencies. A case in point is that for long ods short-wave broadcasting services are now carried out on uencies as high as 26 Mc/s, whereas in 1939 very little use made of frequencies higher than 17 Mc/s and none at all of uencies higher than 21 Mc/s.

nproved use of frequencies has also come from improvement eceivers and in the use of space-diversity and frequency- rsity reception for relaying broadcast programmes.

(4) FREQUENCY PLANNING

he arranging of inter-governmental conferences on the d allocation of the radio-frequency spectrum among the ous services is the responsibility of the International Tele- munication Union, now a specialized agency of the United ions Organization. The task of the I.T.U. is one of ever-increasing difficulty and complexity owing to the continual increase in the number and power of radio-frequency transmissions. The frequency-band allocations currently in force were part of the Radio Regulations agreed at a Conference at Atlantic City in 1947.[16] The range of frequencies then allocated was from 10 kc/s to 10 Gc/s. A revision of this allocation table was made at Geneva in 1959—the new agreement being operative from May, 1961—in which the range of frequencies has been extended to 40 Gc/s. The bands allocated to sound broadcasting and applicable to the United Kingdom, which do not differ greatly from those allocated at Atlantic City, are given in Table 1.

Table 1

Band	*Frequency Range*	*Remarks*
Long-wave	150–285 kc/s (2000–1053 m)	One frequency (200kc/s) available to B.B.C.
Medium-wave	525–1605 kc/s (571–187 m)	12 frequencies for Home, Light and Third Programmes and two for European Services, plus two international common frequencies, of which one is at present in use.
Short-wave	3950–4000 kc/s (75 m band) 5950–6200 kc/s (49 m band) 7100–7300 kc/s (41 m band) 9500–9775 kc/s (31 m band) 11700–11975 kc/s (25 m band) 15100–15450 kc/s (19 m band) 17700–17900 kc/s (16 m band) 21450–21750 kc/s (13 m band) 25600–26100 kc/s (11 m band)	126 frequencies are registered for use by the B.B.C. Overseas and European Services.
Band II (V.H.F.)	87·5–100 Mc/s	40 frequencies at present mainly restricted to 88–95 Mc/s, used for B.B.C. Sound Services.

The assignment of frequencies within the broadcasting bands to individual stations has been confided to special conferences sponsored by the I.T.U. A Short-Wave Broadcasting Conference was held at Mexico City in 1948, when draft plans for one season of a sunspot cycle were prepared. The Conference continued in Florence, and later in Rapallo, but no agreement could be reached because the broadcasting requirements submitted for other periods of the sunspot cycle could not be fitted into the frequency bands available. A considerable amount of short-wave broadcasting outside the allocated bands followed this failure to prepare acceptable plans. For some years prior to the 1959 Geneva Conference special efforts were made to find in-band frequencies for those broadcasting stations working out-of-band. Some progress was made, but the problem still remains.

One important step taken at Atlantic City was the setting up of the International Frequency Registration Board—a panel of 11 experienced radio engineers. Its main task is to draw up and maintain an international list showing all frequency usage throughout the sunspot cycle. While possessing no mandatory powers, the I.F.R.B. makes a technical study of all frequencies submitted for inclusion in the list and can warn and advise administrations about interference likely to result from the introduction of new or changed services. At the Geneva Conference this Board was given the extra task of studying the short-wave schedules of every broadcasting organization in the world for each season, and advising all the countries concerned how best to avoid mutual interference so as to make the best use of the broadcasting bands available. This 'frequency management procedure' may help to lessen the need for the continued use of out-of-band frequencies.

The allocation of long-wave and medium-wave channels to individual stations in the European region was undertaken by

the Conference at Montreux in 1939, following the International Radio Conference of the I.T.U. held at Cairo in 1938. The Montreux Plan was never put into effect because of the war, and most countries in Europe continued to operate during the war years in broad accordance with the Lucerne Plan of 1933. After the war there was growing congestion in these wavebands because the increased importance of broadcasting in the social and political life of European peoples had resulted in the appearance of many new broadcasting stations. Furthermore, technical developments had made possible the use of transmitters of higher power. A new plan was therefore evolved at Copenhagen in 1948, which came into force in March, 1950.[44] In accordance with this plan, the B.B.C. uses one long wavelength and 12 medium wavelengths (plus one international common frequency) for its home services and two wavelengths for its services to Europe. This plan partially succeeded for a time in arresting further deterioration, but a great deal of mutual interference between stations in different countries has since developed, mainly because the plan was not adhered to by all European countries and indeed was not accepted by several of them, including the occupying powers in Germany. As a result of these and other difficulties the number of stations using these bands had increased from some 520 provided for in the plan to about 1 000 in 1959. Moreover, deliberate jamming was introduced by certain countries in these bands as well as in the short wavebands, and this still persists. A serious degree of interference has therefore developed after dark, especially in winter, and this has been the main factor leading to the rapid development of sound broadcasting on very high frequencies in certain countries, notably Germany and Italy, as well as the United Kingdom.

A realization of the importance of planning the v.h.f. broadcasting bands while development in many European countries was at an early stage led to the convening of a European Regional Conference at Stockholm in May and June, 1952. Some 30 European countries were represented at the Conference, many of whom had at that time no immediate plans for developing television or v.h.f. sound broadcasting services, but who wished to see possible future services catered for in an agreed European plan.

The Conference[47] produced plans for the use of v.h.f. Bands I, II and III, and, realizing that these plans could not be too rigid owing to the uncertainty of future developments in most of the countries represented, suggested that a further Conference be held in 1957. (This Conference will be held in 1961.)

(5) TRANSMITTING STATIONS

The following are among the most important developments in the design of transmitting stations:

(*a*) The trend towards higher power and increased hours of transmission emphasized the economic importance of achieving higher efficiency, particularly in the modulation process. After pre-war trials with the Chireix and Doherty systems, and more recently with the ampli-phase system, class-B modulation remains in general use for a.m. transmitters. For f.m. transmitters the efficiency of the modulation stage is less important; this stage does not consume much power because modulation is applied directly to the r.f. drive. The a.c./d.c. conversion efficiency of r.f. stages has been increased also.

(*b*) The need for rotating machinery has been greatly reduced by the use of mercury-arc rectifiers instead of motor-generators for the h.t. supply, by evaporative cooling of valves and by the use of alternating in place of direct current for cathode heating. The application and control of these supply voltages has been made automatic, and this has facil unattended operation.

(*c*) The use of thoriated filaments in transmitting has resulted in increased efficiency and more compact ment.

(*d*) Increased use of air-cooled, instead of water-c valves and more recently of evaporative cooling has f reduced the size of equipment for a given power an further improved reliability and efficiency.

(*e*) Improvements in the design of anti-fading mast rad for medium wavelengths.

(*f*) The development of v.h.f. aerials (including slot a which give a substantial gain in the horizontal plane an also have a directional radiation pattern; these aeria usually carried on the same masts as television aerials.

(*g*) The design of combining circuits for v.h.f. st permitting three or four sound programmes to be radia the same aerial and incorporating filters to reduce r.f. modulation products.

(5.1) The Development of High-Power Transmitters

Prior to 1934 the majority of the high-power transmi use had an output power of about 60 kW, and their c efficiency was no more than 20%. These transmitters the low-power modulated type. By 1937, high-power modulation was coming into general use, and new trans were operating with an output power of about 100 kW. efficiency of these transmitters was somewhat better, be 30% and 40%. In 1954 high-power transmitters of 150 k put using air-cooled valves with thoriated filaments were co and efficiencies of some 40–50% were realized. Some mitters of higher power than this are used for broadca particularly in the long-wave band where the greater non- range makes this worth while. The Light Programme mitter at Droitwich, for example, has an output of 400 k the long-wave Allouis (France) and Moscow transm 500 kW. During the war, the B.B.C. installed an 800 kW wave transmitter consisting of four 200 kW units, two, th four of which could be operated in parallel as desired. present, the *Voice of America* medium-wave transmit Munich is operating with a power of 1 000 kW.

The trend in transmitter design is well exemplified Third Programme and Network Three transmitter at Da which was brought into service in April, 1951. This trans is built in two halves, each capable of an output power of 10 These are normally operated together at somewhat below maximum rating to give an output of 150 kW, the max permitted in the Copenhagen Plan. The combining circu so arranged that, if a fault develops in one half of the trans the programme continues without interruption, thou reduced power, using the other half. In this transmitt cooled valves are used throughout, a new departure in a transmitter of this power. Another feature of this trans is that all the valve filaments are heated by alternating c so that there is no rotating machinery, other than air-b associated with it. A third feature of particular interest the transmitter is designed so that it can be operated f remote point and is, in fact, so operated from the shor transmitter building on another part of the site. High- class-B modulation is used, and an overall efficiency of 45% is obtained.

Considerable reduction in the dimensions of transmitt a given output power has been achieved by the use of air- valves and through the saving of components which follo adoption of single-ended circuits using the inverted-am technique.

number of high-power short-wave transmitters recently red by the B.B.C. use evaporation-cooled valves—an old which has become practicable with the development of ble valves and sub-units. In this system, the heat generated ie valves is transferred to water jackets surrounding them, the resultant steam is taken by convection to a condenser, h converts it once more to water for return to the water ets. The heat recovered by the condenser may be used for ling heating or to provide domestic hot water.

ie use of high-power class-B modulators and high-power -C r.f. amplifiers can make a large reduction in power costs large station compared with a similar installation using low-er class-A modulators and high-power Class-B r.f. amplifiers. example, at a large short-wave station with a total trans-er output power of some 1 500 kW the saving can amount to 000 per annum, or about 25% of the total running cost. her economy has been achieved at high-power stations by use of mercury-arc rectifiers for the h.t. supply in place of ting machines, which results in a gain in efficiency of e 12%.

emote or automatic control of low-power transmitters and stem of automatic monitoring are now in widespread use as eans of economizing in technical staff.[18–20] At the end of), 28 transmitting stations were operated in this manner by B.B.C. Unattended operation of high-power v.h.f. trans-ers has also been introduced at combined v.h.f. and television ions.

(5.2) Aerial Systems

he shortage of frequency channels remains the greatest tacle to the provision of a nation-wide broadcasting service, it is therefore of the utmost importance that the maximum should be made of the frequencies allocated. The coverage igh-power medium-wave transmitters has been extended by gning more efficient anti-fading aerials, the latest of which he centre-fed mast radiator.[21] Top-capacitance loading, sectionalized masts which enable a loading inductance to inserted between sections, are common means of adjusting electrical properties of a mast radiator and thus influencing nti-fading characteristics.

irectional aerial systems for medium wavelengths are now ely used in the United Kingdom either to increase the field-ngth in a particular direction or to reduce it in the direction another transmitter sharing the same channel. Complex al systems are much used for the latter purpose in the United es, while at a station in Germany erected just before the to give a strong signal in the United Kingdom for propa-da purposes, an aerial system comprising eight masts each roximately 394 ft high and two other masts each 492 ft high used. An elaborate aerial system is also used by Radio embourg. This increases the signal strength in this country, consists of three vertical radiators each electrically some-t longer than half a wavelength and suitably phased to duce a gain of 6 dB over that of a half-wave vertical aerial, he required direction.

he synchronization of the carrier frequencies of broadcasting smitters in the United Kingdom has been developed to a degree of precision, and the B.B.C. now operates 55 lium-frequency transmitters in 11 synchronized groups. The of synchronization has led to increased coverage being ined with the limited number of medium-frequency channels cated to this country under the Copenhagen Plan of 1948.

(5.3) Transmitter Drive Equipment

he use of medium-frequency transmitter synchronization necessitated the development of transmitter drive equipment capable of meeting a much closer frequency tolerance than is required by the current international Radio Regulations for broadcasting transmitters. It has also been necessary to develop frequency comparison equipment to permit a high precision of carrier-frequency identity to be maintained. For medium-frequency transmitter synchronization purposes the B.B.C. maintains a carrier-frequency stability and accuracy of $\pm 0 \cdot 025$ c/s, which results in the field-strength variations of the received signal being comparable with the variations due to medium-wave fading during the hours of darkness.

The B.B.C. Light Programme transmission on 200 kc/s is maintained to a higher degree of precision, namely ± 1 part in 10^8, in view of its use in this country and in Europe by a large number of organizations and manufacturers for frequency standardization purposes. For example, in the German Federal Republic the Sudwestfunk organization uses this high-precision 200 kc/s transmission in conjunction with frequency synthesizing equipment as the drive for broadcasting transmitters in a syn-chronized group.

The precision transmitter drive equipment now used by the B.B.C. has been evolved over a period of 25 years commencing with tuning-fork oscillators which maintained a carrier-frequency stability of the order of one or two parts in 10^6 for short periods of time. This tuning-fork drive equipment occupied four apparatus racks 9 ft in height, and each set of drive equipment cost approximately £3 000 in 1937. Since 1938 crystal drives have been increasingly used and new designs developed. The current B.B.C. medium-frequency crystal drive occupies only 4½ in of apparatus bay space, and cost, in 1960, only £150. Its frequency stability and accuracy, however, are 50 times better than those of the early tuning-fork drive equipment.

In the h.f. broadcasting field, transmitter drives have developed from the simple crystal oscillators used at the original Empire station in 1932. In 1939, high-stability variable-frequency drives were developed to meet the constantly changing carrier-frequency requirements of the service, and at the present time, use is made as necessary of variable-frequency drives, precision crystal drives or frequency synthesizers, the latter being used for the synchronization of h.f. transmitters at stations geo-graphically remote from each other. This latter application embodies the use of the B.B.C. 200 kc/s transmission as a high-precision reference frequency, and by synchronizing h.f. trans-mitters radiating on widely separated bearings, the usage of a limited number of h.f. channels is considerably increased.

(5.4) V.H.F. Transmitters

The introduction of v.h.f. sound broadcasting has resulted in the development of transmitters for operation in the frequency range 87·5–100 Mc/s having various output powers from 2 to 20 kW. These transmitters have air-cooled valves throughout, and considerable use is made in the high-power stages of single-ended earthed-grid circuits with coaxial-line tuning elements. To give maximum reliability, these transmitters are operated in pairs; for example, at the highest-power stations, two 10 kW transmitters are used for each programme to produce a trans-mitter output power of 20 kW. This power is fed to a high-gain slot aerial[46] giving an e.r.p. of 120 kW. The aerial consists of a metal cylinder some 6½ ft in diameter and 110 ft long which has cut in it 32 vertical slots arranged in eight tiers. Each tier has four slots spaced at intervals of 90° round the surface, each slot being about 8 ft high and 1 ft wide. To guard against possible aerial breakdown, the aerial system is divided into two halves which are connected by separate feeders to transmitter com-bining circuits situated inside the station building. The overall efficiency of a 20 kW f.m. transmitter is about 30%.

Two types of transmitter, made by different manufacturers, have been used at these stations, the main technical difference between the two being the method of modulation. In one case a directly-modulated quartz crystal is used.[22] In the other, an *LC* oscillator is modulated by a reactance valve, the centre frequency being controlled by a quartz-crystal oscillator.[23]

In order to extend the v.h.f. coverage and make the maximum use of the available frequencies, the B.B.C. has recently designed a low-power 'translator' equipment for use at satellite transmitting stations. This equipment receives the programme from an existing station and re-broadcasts it on another frequency. Considerable use is being made of transistors in this equipment, which is designed for unattended operation.

The design of feeders for v.h.f. presents difficulties because of the relatively high power (up to 50 kW) that has to be carried at these frequencies over distances up to 1 000 ft with minimum loss, and also because of the need for precise and stable impedance matching. Either rigid coaxial tubes made up from 12 ft lengths, with expansion joints every 150 ft, or flexible cables in continuous lengths up to 600 ft are used.

(6) STUDIOS AND STUDIO CENTRES

(6.1) Studio Design

Many of the improvements and advances in studio design have come about as a result of developments in equipment and of changes in presentation methods and production techniques; a general improvement in studio acoustics has resulted from extensive studies in the laboratory based on new methods of acoustic measurement, from experience in the design of studios and from the availability of new materials. The tendency towards the use for dramatic productions of large studios with several microphones and locally modified acoustics, instead of several smaller studios each with marked individual acoustic properties of its own, linked by a dramatic control panel, has led to changes in studio design. The emphasis has shifted towards medium-size general-purpose studios, each containing two or more sections having radically different acoustic treatment. Often the layout of a studio is arranged so that opposite walls are not parallel, and in the larger studios the ceiling is broken up or coffered.

Talks and discussion studios have become somewhat larger than those used before 1939. Great importance, from the point of view of speech clarity, is attached to having adequate sound absorption at the lower frequencies. The discussion type of programme with four to six speakers has largely replaced the single speaker. This has had considerable repercussions on the type of microphones and, above all, on the studio acoustics. A specially-designed circular acoustic table, used with a cardioid microphone, has been found necessary, and the desirable acoustics are somewhat deader (reverberation time, 0·3 sec) than was formerly the case for speech studios.

In the last few years it has become the practice to adopt a multi-microphone technique for light-entertainment programmes, often using as many microphones as there are instruments in the orchestra. In addition, artificial reverberation methods are fully exploited and hence the actual studio acoustic requirements are somewhat wider. Reverberation times of 1·0–1·6 sec are considered satisfactory.

Very considerable research into the desirable acoustic properties of concert halls and large orchestral studios has been undertaken, and the results in the form of porous and resonant absorbers and architectural features designed to scatter reflected sound have been successfully applied in many studios.[24] Notable examples are the large orchestral studio at Maida Vale and the rebuilt studio premises at Swansea.

(6.1.1) Artificial Reverberation.

From the earliest days of broadcasting until a few years the standard method of obtaining artificial reverberation by the use of a small reverberation chamber or 'echo' r Several attempts, only with moderate success, have been r to replace the echo room by using a series of spaced heads magnetic drum. This method suffered from quite a numb faults, in particular its failure to cope with percussiv impulsive sounds and longer reverberation-time requirem e.g. cave effects, etc. More recently greater success has achieved by the use of a thin, freely suspended 6 ft × 3 ft plate. The time of reverberation can be varied from to 1 sec by moving a damping membrane closer to the Because of their small size, compared with that of an chamber, and the ability to vary readily the reverberation over a wide range, artificial reverberation plates are increasingly used on all types of programmes, especially music and variety. They are also very useful for correctin transmission the acoustic shortcomings of public hall theatres, especially opera transmissions.

The very great expansion in the daily programme outp the B.B.C. has resulted in an increase in the number of s broadcasting studios from 95 at the beginning of 1939 at the end of 1960. Of these, 35 are used for the ext services.

In the last ten years a chain of unattended studios has established throughout the country at various locations, s menting the more extensively equipped regional centres sub-centres. These unattended studios serve news and cality requirements, enabling reports to be broadcast wit the necessity for the speakers journeying to the main cen which may well be some 40 or 50 miles away. The equip and technical arrangements are such that no engineer nee present—hence the term 'unattended'. The studios are talks variety, about 20 ft × 16 ft in size, with an adjacent r housing the control equipment. Arrangements have been so that the studio can also be used for more elaborate t missions with a producer and engineer present in the adj control room. There are 12 of these studios in servic present and five more in the course of construction.

The use of general-purpose studios and the development c self-contained studio unit in the B.B.C., arising partly from stringencies of war conditions, have several advantages— ticularly in relation to rehearsals. Before the war, the ce control room at a studio centre handled both live transmis and rehearsals, which added considerably to the operational In later designs, all the technical facilities needed for rehea and most of those needed for transmission are provided i local studio control cubicle. For rehearsals, the whole o programme-originating equipment can feed into the local speaker without going through the central control room.

(6.2) Studio Equipment

The war brought many problems for those associated wit design of audio-frequency equipment, not the least of w were the impossibility of continuing with certain designs o to the lack of materials, and the necessity of equipping a number of new studio premises at very short notice. Thi to the introduction of 'standard bays' based on the OBA/8 of outside broadcast equipment and other units of simp design, which could be produced in quantity, assembled wired in the workshops and connected in position with minimum of delay. New equipment designed by B. engineers is radically different from that used before and d the war.[25] Studio equipment introduced in 1947 and kn as type A consists essentially of two main items—a co

: and an apparatus cabinet. The former is installed in the io control cubicle and houses faders, talkback microphones, icial-reverberation facilities, keys for bringing in spare lifiers, programme meter, telephone instruments and similar pment. The apparatus cabinet houses all the amplifiers, er units, relay and other technical apparatus and need not ssarily be situated in the control cubicle. The amplifiers in cabinet are connected to the permanent wiring by plugs and ets, and the relay assemblies by Post Office type springjacks. se can thus be readily removed for maintenance or repair.

new type of studio equipment known as type B has e been developed and was introduced into service in 1954. ; is less costly than type A and provides the many more lities and increased flexibility demanded by refinements in luction techniques. The new equipment has been designed the unit principle and comprises three standard amplifiers 1 the necessary mains supply units, standard panels containing ers, cue keys, indicators and similar apparatus, all of which be assembled in numbers suitable to the size of the studio the production facilities required.

he amplifiers used in the control rooms were all battery-rated up to the beginning of the war. These have since been erseded by mains-operated types, fed in groups from mains ts, which themselves have steadily been reduced in size and, re important, in cost. Amplifiers incorporating transistors e been designed and are being brought increasingly into vice.

(6.3) Continuity Suite

he technique of assembling programme items and passing m to the transmitting stations has also undergone a radical nge in the period under review. Whereas before the war it s customary for the central control room to handle all pro-mme switching, this work has now been concentrated in tinuity suites, one of which is associated with each of the in programme services.[26] A continuity suite comprises a all studio and a miniature control room. These two areas connected electrically and also visually by means of a sound-of window. All programme changes are made in the con-uity control room by a technical operator, while linking and announcements, 'fill-ups' and similar material are handled the continuity announcer.

n the event of a breakdown or other emergency the announcer on the spot to explain the situation to listeners within a few onds of the occurrence. This method of working not only ves the central control room free to concentrate on the eral operation of the B.B.C. services, but ensures smoother l more closely knit programme presentation.

(6.4) Central Control Rooms

t the beginning of the period under review, all control rooms B.B.C. studio centres used remotely-controlled relay switching the routing of programmes. The rapid war-time growth in number of sources and destinations which had to be switched ild have rendered the continued use of this system imprac-ble because of the sheer bulk of the equipment, even if the nendous number of relays needed had been available. It ; therefore necessary to revert temporarily to a simple plug-l-jack system of interconnection. This elementary method ow being replaced by a greatly improved arrangement using tor uniselector switching which provides for the simultaneous ction, by simple code, of programme circuits with the asso-ed cue, control and signalling circuits.[27] Very large numbers ources and destinations can thereby be controlled from and cated at a single control position without operational com-ity. The latest example of this technique is the new control room at Bush House for the External Services where the push-button switching can connect 150 sources to 130 destinations. Control rooms at the regional centres are also being modernized and re-equipped with motor uniselectors for programme, moni-toring and cordless engineering telephone-exchange switching.

(6.5) Automatic Programme Routing and Monitoring Equipment

The External Services* require frequent changes in the pro-gramme connected to particular transmitters, and these connec-tions may need to be altered at intervals of 15 min throughout the 24 hours in accordance with a specially prepared schedule. The schedule is repeated every 24 hours, and is only changed at intervals of, perhaps, some months.

The transmitters themselves are distributed among a number of stations in different parts of the country, each station being connected by a group of lines to the Bush House control room. Ideal conditions therefore exist for the use of automatic switching equipment at Bush House to connect programme sources to the appropriate lines and for similar equipment at transmitting stations to connect the incoming lines to the appropriate trans-mitters.[27, 28] Automatic time-controlled switching equipment has been specially designed by the B.B.C. for these purposes. The automatic equipment can, of course, be overridden by manual control if necessary. At transmitting stations, auto-matic apparatus is also arranged to sample the programme incoming on a particular line and to follow this by sampling the outputs of all transmitters connected to that line. These samples, each of 5 sec duration, are connected in turn to a loudspeaker and are thus presented as a continuous series of excerpts for aural monitoring.

The shortage of skilled staff and the need for further economy has led the B.B.C. to develop automatic equipment for the monitoring both of the lines between studio premises and trans-mitters and of transmitters themselves.[29] The increasing use of equipment of this type has enabled improvements to be made in the coverage of the domestic programmes with a more economical use of technical manpower than would have otherwise been practicable, at the same time relieving technical staff from much routine work and making them available for more interesting tasks.

(7) MICROPHONES AND LOUDSPEAKERS

Developments in microphones have included substantial improvements in, and a reduction in the size of, the ribbon microphone, and the re-appearance of the electrostatic micro-phone (previously abandoned because of the unreliability of earlier pre-war models).

Advances in design and performance have resulted partly from increased understanding of the basic principles and partly from the availability of improved materials, particularly magnetic materials.[30]

In 1939, the standard studio microphone was the B.B.C.–Marconi ribbon microphone type A. This was modified in 1943 by the addition of a new type of ribbon and balanced wiring, and designed type AXB. In 1944 the permanent-magnet system was altered and a Ticonal magnet substituted. Micro-phones of this type were known as type AXBT and had improved sensitivity.

The demand for smaller and lighter high-quality microphones led to considerable research and to the trial of many different types. One new type introduced into service in 1952 is known as the PGS† ribbon microphone developed by the B.B.C. Research Department and manufactured by Standard Tele-phones and Cables, Ltd. It is considerably smaller and has

* The present output of the External Services comprises ten programme services in 39 languages totalling 84 hours per day.
† British Patent Nos. 738664 and 742006.

improved characteristics, particularly at the higher frequencies, than the type AXBT and weighs only one-third as much.[31] This microphone represents a considerable improvement in quality, particularly in its response at the higher audio frequencies, and is less obtrusive than the type AXBT. It is known as type 4038.

However, there is considerable demand for microphones possessing characteristics other than that of the figure-of-eight (as in the ribbon microphone)—in particular, we require cardioid and omnidirectional characteristics. Hence it has been necessary over the last 15 years to make increasing use of electrostatic microphones, which, because of the small size of their working parts, can readily be designed to have any, or in some cases all, of the desirable directional characteristics without appreciable frequency distortion. In particular, the cardioid type is essential for outside-broadcast use in the footlights of theatres and opera houses in order to obtain the maximum of stage pick-up without the orchestral pick-up on the back of the microphone. The omnidirectional type is very useful in places where the reverberation time is too short for ideal results.

For use in conditions of high ambient noise, e.g. when a commentary on a sporting event is broadcast, a close-talking pressure-gradient microphone has been designed by the B.B.C. Research Department. The original design was introduced in 1937 and became known as the lip microphone type L1; an improved version, type L2,* was produced in 1951 and was, until recently, the only high-quality 'noise-cancelling' microphone of its kind in the world.[45] A microphone, based on the B.B.C. design, is now manufactured by Standard Telephones and Cables, Ltd., and designated type 4104B or C.

The lip microphone, when held close to the commentator's mouth, gives a high degree of discrimination between the wanted speech and the ambient noise. When so used, however, speech tends to sound unnatural, and equalization of the frequency characteristic, both by acoustic damping screens and by electrical networks, has been introduced to compensate for this as far as possible.

Advantage is taken of commercial products where microphones having special characteristics are required and are available, for example directional properties or robustness, or where there is need for a microphone to be held in the hand or worn on the person. The B.B.C. maintains an acoustics laboratory with advanced facilities for the calibration and testing of microphones.

The critical aural monitoring of programmes, which is carried out at several points in the broadcasting chain, demands a high-quality loudspeaker used under suitable conditions to show up any unwanted noises, distortion or other technical faults. It must also reproduce sound with a high degree of fidelity so as to enable a balance to be obtained that is acceptable when the programme is heard over a wide variety of reproducing equipments and not merely over other loudspeakers of the same type.

The design of a high-quality loudspeaker is a frustrating task for the engineer, since the characteristics required have so far defied attempts to reduce them to a precise specification. The performance of the final product has to be assessed by ear, but unless the subjective judgment is arrived at under carefully controlled conditions, it is easy to obtain completely contradictory results. To provide itself with loudspeakers of the necessary high standard of performance the B.B.C. has, in addition to laboratory tests, made direct comparisons between the reproduced sound from a loudspeaker and the original sound in the studio, employing for the purpose its own trained and experienced staff. In this way, the performance of new commercial designs is rigorously assessed. The B.B.C. also designs its own loudspeaker assemblies, by combining units from different manufacturers.

* British Patent No. 737096.

A loudspeaker designed for monitoring purposes b B.B.C.[32] uses a commercial 15 in low-frequency unit moun a vented cabinet in conjunction with high-frequency unit suitable cross-over networks. A smaller and lighter ad of this design has been produced for use in outside broa Tests have also been conducted with modern commercial ve of a wide-range electrostatic loudspeaker.

The built-in 'corner' loudspeaker possess several advan but under operational conditions it is seldom possible t this type.

(8) OUTSIDE BROADCASTS

In 1939, the standard B.B.C. equipment for outside broa consisted of the OBA/8 amplifier and its associated appa Little progress was possible in the design of new equi during the war, although the requirements for outside-bro operation had become more complex. In order to provi additional facilities required in post-war years, unsuitable by modern standards—rather cumbersome equipment h be pressed into service, and the number of units to be trans and connected on site had become unwieldy.

The OBA/9 amplifier and its auxiliaries were develo provide equipment as small and light as possible w sacrificing reliability, yet having a general technical perfo similar to, or better than, the OBA/8. Ease of transpo if necessary, dismantling, were considered important provision for carrying subsidiary items such as micro cables, cue lights, telephones, etc., was made.

The resulting equipment is mounted in a stack on a porter's trolley, which facilitates transport. It consists units of five different types: a 4-channel mixer, a micro amplifier and programme meter—of which two are inclu each set—a distribution and general control unit, a loudsp and isolating amplifier, and a power supply unit. The equi can be operated from the mains supply or from dry ba contained in the power supply unit. The equipment designed that these six units can be operated without disma or alternatively they may be removed from the trolley an the addition of a suitable number of similar units, exten deal with the more complex programmes. Mounted at th of the trolley are three drums each containing 150 ft of sc microphone cable, while two additional items—a portable speaker and a spares and components box—are trans separately.

In designing this equipment full advantage has been ta the availability of small valves and other components so keep its size and weight to a minimum. A transistorized v of the OBA/9 amplifier has now been produced and is in r service in the sound and television services. Its lightnes power consumption and freedom from microphony ar ticularly valuable.

Both OBA/8 and OBA/9 equipments were used in pro sound facilities during the Coronation broadcasts in 19 which 1 300 additional sound circuits were required to li 11 temporary control rooms and more than 80 control po with the permanent B.B.C. control rooms in the London The five main sound networks covered broadcasts and reco by more than 100 commentators speaking in 42 languages

Equipment of the same types has been used for the tours of Canada, South Africa and Southern Rhodesi Australasia.

A new development in the outside-broadcast field for programmes is the suitcase outside-broadcast apparatus, is designed for operation by a commentator or reporter w the need for an engineer. The whole apparatus is contai a small suitcase measuring $15\frac{1}{2} \times 9\frac{1}{2} \times 5\frac{3}{4}$ in, and the

ht is 15 lb. Transistors are used in the microphone ampli- ınd also in the very small radio receiver, by means of which ramme cues may be obtained when required from the usual C. transmissions. A crystal microphone is provided, and hone ringing and speaking facilities are included. The user nects his equipment to the appropriate Post Office line circuit nserting a single plug into a socket provided at each outside dcast point.

nother B.B.C. development, used mainly in outside broad- s, is a radio microphone which was designed to relieve mentators of the encumbrance of a microphone with a ing lead. This was introduced in July, 1955, and is used ooth the sound and television services. It consists of a iaturized v.h.f. f.m. transmitter and a battery pack each roximately the size of a packet of 20 cigarettes, associated a miniature microphone which can be worn on the person. aerial consists of a few feet of wire concealed in the user's hing. The transmitter operates on frequencies in Band I has an r.f. output of about $\frac{1}{4}$ W. It is believed to be the llest high-quality transmitter of its kind at present available this particular purpose. The range of the transmitter ends upon conditions, but under favourable circumstances exceed half a mile.

ecent additions to the fleet of special outside-broadcast icles have been two versions of a mobile studio and control m fully equipped to carry out large-scale programmes. In more recent version the control-room equipment can handle otal of 23 programme sources.[33] Two of these are micro- nes in the vehicle's own studio, one is the output of a bank our disc-reproducing turntables and four are the outputs of f. f.m. receivers bringing programme items from transmitters ied by commentators in the field.

ound programmes are usually brought from the outside- adcast point to a convenient point on the main network by ins of lines temporarily hired from the Post Office and tested equalized by the B.B.C. Where the outside-broadcast point nobile, or for any other reason, suitable lines are not available, io links are used.

(9) SIMULTANEOUS-BROADCASTING LINES

he permanent network of lines connecting the studios with transmitters, both for conveying the programme and for nmunication purposes, is supplied and maintained by the t Office. B.B.C. engineers co-operate with those of the t Office in setting up the programme lines to the high standard uired and carry out routine tests to ensure that the specified nical characteristics are maintained. The standards aimed are at least equal to those laid down by the C.C.I.T.T. ernational Telegraph and Telephone Consultative Commit- . Equalizers are used at B.B.C. centres to compensate for residual attenuation/frequency distortion; for long pro- nme circuits it has been found necessary to introduce perature compensation into the equalization.

he expansion of the B.B.C.'s various services has demanded roved communication facilities. The comprehensive com- nication system that has been built up now provides both r-office telephone communication and engineering control lities between all the regional centres and transmitting ions throughout the network; it also provides teleprinter lities between London and the main B.B.C. centres. Most he communication circuits are provided by 3-channel or annel carrier systems each employing two Post Office music uits. In some cases the music circuits can be used for pro- nme transmission at certain times of the day, and for inter- e communication at other times. The teleprinter channels, which are derived by the use of splitting filters, employ frequencies between the speech channels. The B.B.C. also uses facsimile, telegraph and picture equipment in its television news service.

The B.B.C. now rents from the Post Office about 25 000 miles of programme circuits for sound, including those used for the sound component of the television programme, and 7 000 miles of control and communication circuits.

The programme networks comprise a distribution system for carrying each of the programmes from the continuity centre to the transmitting stations and a contribution network bringing programmes from remote parts of the system to the continuity centre. The main networks are set up in programme chains, i.e. Home, Light, Third and Television Sound, and they are normally routine tested each week for attenuation/frequency characteristic, harmonic distortion and noise, the test signals being originated at the London end and readings being taken manually at intermediate studio centres and transmitters. Contribution circuits are similarly tested in chains. This method, however, does mean that a number of engineers are engaged simultaneously on the tests, which take about 30 min for each chain. Automatic sending apparatus has now been introduced for these tests on the main programme distribution chains, thereby reducing the time to 5–6 min. Apparatus has also been designed to reduce this time still further to 3–4 min with the synchronized automatic adjustment of receiving apparatus to meet the requirements of each test and the recording of results without manual assistance. It is intended to use this apparatus also for the adjustment of 'topping up' equalizers on long chains of three or four links so as to 'iron out' the accumulated variations of the attenuation/frequency characteristics. These equalizers use Bode-type sections to correct the frequency characteristic in three parts of the frequency range, i.e. 60 c/s–500 c/s, 500 c/s–2 kc/s, 2 kc/s–8 kc/s, by approximately ± 4 dB in each part. Equalizers of types similar to these have been in service for a number of years for the correction of 'occasional order' circuits, which can be successfully and rapidly trimmed by testing at four main frequencies.

Other developments during the period have been the increased use of carrier circuits, and the development of special techniques, such as a split-band system for use where the available lines have insufficient bandwidth.

(10) USE OF REBROADCAST RECEIVERS IN PLACE OF LINES

With the development of the v.h.f. f.m. system, increasing use has been made of the practice of feeding transmitters by direct pick-up from a v.h.f. f.m. station in place of a line. By this means, economy in line costs is effected, and where existing lines have been inferior or liable to interruption, this change-over has made possible an improvement in the quality and reliability of programme distribution. B.B.C. design v.h.f. f.m. rebroadcast receivers are provided at the transmitting station which is to be fed in this way and are tuned to a convenient station radiating the required programmes. The originating station becomes the parent station, and the receiving station becomes the satellite. The satellite station uses the output of the receiver as the source of programme for its transmitters in place of the normal line source, and may, in its turn, become a parent station for a further group of satellites.

It has been possible to extend this system to unattended and semi-attended transmitting stations by the use of a system of 20 kc/s tone monitoring. Two complete chains, one normal and one reserve, are provided at these stations, each comprising a rebroadcast receiver, programme input equipment and transmitters. The parent station radiates a 20 kc/s tone at a fixed

level superimposed on the programme. Tuned amplifier-detectors are provided at the satellite stations which detect the tone and initiate the switching as required for the particular station. In general, if the 20 kc/s tone is not detected at the appropriate level on the normal programme feed circuit, the receivers are changed over to the reserve chain. If the tone is not detected on this either, then the transmitters may be connected to other sources of programme (long-wave or medium-wave rebroadcast receivers or lines) or may be shut down. In some cases, the amplifier-detector is also used to switch on the transmitters.

Radio reception of the transmissions from v.h.f. f.m. stations for programme distribution was first developed for use with certain low-power medium-wave stations and was then extensively introduced into the programme distribution network for v.h.f. f.m. transmitting stations. It is likely to be further extended to other stations.

(11) SHORT-WAVE BROADCASTING

In 1938 there was comparatively little short-wave broadcasting outside that provided by the United Kingdom, Germany, the Soviet Union and the United States. During the period under review the number of short-wave transmitters in the world has increased enormously. The power of the transmitters used has also increased from a general level of 5–10 kW in 1938 to 50–100 kW and above.

In 1938 the B.B.C. operated six short-wave transmitters with a total output power of 170 kW. At its war-time peak, in February, 1944, this output had grown to 2870 kW from 43 transmitters. The B.B.C. is at present using 39 high-power transmitters, two of which are installed at Tebrau (Singapore) and are used primarily to rebroadcast the B.B.C. Far Eastern and Eastern Services to the appropriate Asian countries.

The short-wave transmitters have been modernized and improved, where possible, in efficiency by the use of modern valves, improved modulation techniques and measures taken to increase the speed of wave-changing. Plans have been prepared to replace some of the older transmitters, and work is already in progress.

The design of transmitting aerials and feeders has been the subject of intensive study, and high-gain directional aerials are now in general use. These aerials consist of stacked horizontal dipoles—a form which has been found to be generally the most efficient for this purpose.[34, 35] At a large transmitting station a formidable number of aerials is needed to cover the many areas to be served and the number of frequencies required, and a great deal of land is required on which to erect them. Considerable economy and flexibility can be effected by the use of aerials with reversible directivity, by electrical slewing of the main lobe in the horizontal direction, and by comprehensive transmitter-to-aerial switching arrangements.

Owing to the shortage of frequencies, extensive use has been made of synchronized transmitters; the technique has been developed to include transmitters at stations in different parts of the country. Experience has shown this to be a satisfactory method of operation for serving two or more areas with the same programme on the same frequency so long as the aerial patterns of the transmitters do not overlap.

The general increase in the power of short-wave transmitters and improvements in receiver design in regard to sensitivity, noise factor and a.g.c. action has greatly improved the reliability of short-wave services. The congestion of the short-wave bands and the resulting interference has, however, resulted in serious difficulties, and has led the B.B.C. to take steps to increase the effective radiated power of some of its longer-range services. It has been found possible to operate two 75 kW short- transmitters in parallel, and by using them in conjunction an aerial having a gain of 23 dB, referred to a half-wave d in free space, to produce an e.r.p. of 30 MW over a small ar

Much progress has been made in understanding the behav of the ionosphere, leading to improved forecasting of optimum frequencies for short-wave circuits. There has be great increase both in the number of measurements of ionosp conditions and in the number of stations throughout the w at which they are made. A well-organized system for exchange and correlation of data has thus been established co-ordinating authority in the United Kingdom is the D.S A large number of special observations of phenomena affe radio propagation were made during the International physical Year; the B.B.C.'s Receiving and Measuring Stati Tatsfield was chosen as one of the centres for obse ionospheric and tropospheric propagation conditions. (of interest to note that the first observations in Great Br of the signals from the first Russian earth satellite were rec at Tatsfield in the early hours of 5th October, 1958.) now confirmed that the 11-year solar cycle passed throug minimum phase in April–May, 1954, and its maximum February–March, 1958. During the peak solar activity highest frequency bands (21 and 26 Mc/s) available for sh wave broadcasting could be used to a greater extent, thus tially relieving the congestion in the lower frequency bands.

The conditions of operation of a short-wave service th virtually world-wide in its coverage demand the use at diff times of day of a number of different frequencies. All o of those frequencies may be changed *en bloc* when a m alteration in the operating schedule is made to take accou seasonal propagation changes. At the time of writing, 12 quencies are registered for use in the B.B.C.'s External Serv and it would be impracticable to install this number of grade crystal drives at each short-wave station. The B. has therefore developed variable-frequency drive equipmen exceptionally good frequency stability and high rese accuracy.[36] An essential part of this equipment is a frequ monitor which provides, from a local frequency sub-stand standardizing frequencies spaced at 5 kc/s intervals throug the carrier-frequency range of 2·8–22·4 Mc/s. The approp one of these frequencies is compared with that of the transm which is adjusted to zero beat. It should be noted that frequency monitor is not a measuring device, but an ai adjusting frequencies already known to an accuracy of, 50 parts in 10^6 (from the dial setting of the drive equipm to an accuracy within a few parts in 10^7.

The international frequency regulations laid down by Atlantic City Convention in 1947 allow a tolerance of ±3 per Mc/s for all broadcasting transmitters operating on quencies from 4 Mc/s to 500 Mc/s. The B.B.C. norm achieves a frequency tolerance better than ±15 parts in 10 the short-wave bands. Under the new Radio Regula adopted at Geneva in 1959, the tolerance for the 4–29·7 Mc/s will be reduced to ±15 c/s per Mc/s for transmitters installed after 1st January, 1964, and for all tr mitters after 1st January, 1966. Changes have also been in the tolerances for the higher-frequency bands.

(12) B.B.C. MONITORING SERVICE

The B.B.C. Monitoring Service came into being shortly be the war when a need arose to intercept and summarize items broadcast by foreign stations. Initially, monitoring introduced on a small scale in London, but in 1939 the se was transferred to Evesham and put on a professional basis

chnically, the organization was planned in two parts. For tion of the stronger signals an amplified aerial system was lled, covering the frequency range 100 kc/s–27 Mc/s, in h some 60 domestic-type receivers were fed via r.f. amplifiers a number of omnidirectional aerials. For weaker signals, nunication-type receivers were used, connected to individual tional aerials. This scheme suffered from a number of lvantages among which were:

i) Strong signals from local transmitters produced cross-modula- n and inter-modulation effects in the r.f. aerial amplifiers.
ii) The limited area of land available restricted the erection of ectional aerials.
iii) The distance from London (100 miles) was too great for isfactory operation of the service.

site free from these disadvantages was sought, and in 1942 a ble one was found at Caversham, near Reading. The same, plan was adopted, strong signals being intercepted by gn-language monitors operating their own individual vers, while weak and difficult signals were received by neers with special equipment at a remote and electrically site at Crowsley Park some $3\frac{1}{2}$ miles from the main centre.[37]

Caversham, an amplified aerial system was installed which isted of omnidirectional aerials, each feeding an r.f. signal lifier having a frequency range of 2 : 1 and a gain of not than 25 dB. The r.f. spectrum of 100 kc/s–27 Mc/s was ivided into seven octaves and one near-octave. The out- from the octave amplifiers were combined in two groups of and one group of two, and connected to three coaxial es each running to a special distribution transformer. n transformer was capable of feeding up to 60 receivers, ated by the monitors who selected the output from the opriate transformer by means of a switch at the receiver t. This scheme has the advantage that the first-order, and efore strongest, combination terms produced by the octave lifiers necessarily fell outside the frequency range of the lifier from which they originated and within the gaps between ves carried by each cable. It also has the advantage of iding an attenuation of as much as 41·6 dB between receivers.

t Crowsley Park, a comprehensive system of directional ls was installed including five rhombic aerials and 12 rage long-wire types, in addition to 30 long-wire semi- cal aerials. A flexible system was installed in which aerials plugged by flexible cords to any desired octave amplifier. nty-seven communication-type receivers were installed, and a.f. outputs were connected by line to Caversham through ıg-and-jack board.

ost-war developments have included[38]

(a) The replacement of the original octave amplifier system at owsley Park by six push-pull wideband amplifiers having a high gree of linearity and each covering the frequency range 100 kc/s– Mc/s and having ten r.f. outputs.
(b) The substitution of the original American communication ceivers by British receivers of improved performance.
(c) The use of bi-directional rhombic aerials.
(d) The use of very-long-wave convertors in which Morse signals the frequency range 15–150 kc/s are made to key an r.f. oscillator, ıs producing an output within the r.f. range of the normal receivers.

ne main listening room at Caversham has been almost pletely re-equipped in recent years. There are now 40 nal monitoring positions, each equipped with a modern sh communication-type receiver. Remotely-started plastic- recording machines are installed, so that items can be rded for subsequent replay on transcription machines. The rding machines are arranged in groups, and electrical king' facilities are provided at each monitoring position, so any recorder which is not already booked or in use may be ved for instant use when required. Two further positions are specially equipped for the reception of Hellschreiber transmissions, and there is also a remotely-operated magnetic-tape recorder for use when better recorded quality is required than is obtainable from the plastic-belt machines.

The programme lines from Crowsley Park are connected at Caversham to the appropriate monitoring positions by switching at a supervisory control position in the listening room. Here a special console designed by the B.B.C. has been installed, which provides such facilities as pushbutton monitoring-line routing, supervisory monitoring of all monitors' listening positions, and two communication-type receivers for use by the supervisor.

The monitors now cover broadcasts from more than 35 countries in nearly as many languages. From the considerable total intake, the monitoring news bureau selects and processes news and other urgent information for transmission by teleprinter to the B.B.C. news departments and to the Foreign Office. Part of this service is also supplied to subscribing news agencies. A number of publications are produced daily or at less frequent intervals, chiefly for Government Departments; they are available also to other organizations on a subscription basis.

(13) SOUND RECORDING

The pre-war trend towards the increased use of recorded programmes was greatly accelerated by war-time conditions, when it became necessary to record programmes at convenient times for transmission at all hours of the day and to record programmes under conditions not liable to interference from enemy action. A large increase in recording equipment was therefore required.

Prior to the period under review, three sound recording systems were in use by the B.B.C., the Marconi–Stille steel-tape magnetic system, the Philips–Miller film system, and the M.S.S. direct disc recording system. During war time it was decided to obtain a large number of disc recording machines for direct recording on cellulose-coated discs, which can be played back immediately without processing. In addition to static recorders for use at studio centres, portable models using similar discs were manufactured for use by war reporters and for other special purposes. The performance of all these recorders was adequate for the immediate purpose, but it was realized that direct disc recording was likely to become a permanent feature of the broadcasting service, and the B.B.C. therefore undertook the design of a permanently-installed disc recording machine of improved performance.[39] This became known as the type D recording and reproducing equipment, and was brought into service early in 1945.

A great increase in the use of recorded programmes took place in countries engaged in the war, and especially in Germany; towards the end of the war, almost the whole of the German home programmes were recorded. The Germans concentrated on a magnetic recorder known as the Magnetophon, in which the recording medium was a plastic tape impregnated or coated with a magnetic powder. This was highly developed, and extremely good quality was obtained by the use of high-frequency erasing and a high-frequency biasing system. Apart from its high quality, this equipment had the advantage of a playing time of more than 20 min for a single reel of tape, so that a long recording time could be accommodated in a very small volume.

Magnetic-tape recording has now been developed to a high degree of perfection and has come into world-wide use. The speed at which the tape runs past the recording and reproducing heads is of great importance and obviously needs to be standardized if tape recordings are to be interchangeable both within the B.B.C. and with other broadcasting organizations. For

recordings of first-class quality, a tape speed of 15 in/s is used, but equipment is also available giving speeds of $7\frac{1}{2}$ and $3\frac{3}{4}$ in/s. A $\frac{1}{4}$ in-wide plastic tape is used, coated with iron oxide.

International agreement has been reached on the standardization of the essential characteristics of tape and disc recording and reproducing equipment necessary for the international exchange of programmes.

During recent years the B.B.C. has expanded very greatly its facilities for recording programmes on magnetic tape. The major part of the load is now carried by this system and the use of disc recording is decreasing rapidly except for such special purposes as the Transcription Service and for archives. Mobile disc recording has given way entirely to tape. The present proportion is approximately 80% tape and 20% disc.

In addition to the permanently installed recording channels at B.B.C. studio centres, mobile tape-recording equipment has been produced and installed in saloon cars, and there are also portable recorders. The so-called midget recorders are of particular interest—they measure $15\frac{1}{4}$ in × $7\frac{1}{2}$ in × $8\frac{1}{4}$ in and weigh only $14\frac{1}{2}$ lb including batteries. These recorders, which use a tape speed of $7\frac{1}{2}$ in/s, are suitable for recording speech and are supplied to B.B.C. foreign news correspondents oversea, as well as to B.B.C. centres throughout the United Kingdom. The electronic equipment used in these portable recorders has recently been redesigned by the B.B.C., the valves being replaced by transistors. The space saved has been used to accommodate a loudspeaker with its associated transistorized amplifier.

There have been two recent developments of importance brought about by the sheer number of recording channels which have become essential: one is the establishment of separate recording and reproducing rooms where numerous machines are grouped together, thus saving both space and operating staff. These machines are arranged for remote operation, so that once they are set up they can be started or stopped by pressing a button on the control panel of any studio equipped with this facility.

The second development is the trolley-mounted tape machine which can be easily transported to the area where recording or reproduction is required. This has technical characteristics identical with the rack-mounted equipment.

The introduction of fine-groove (long-playing) discs for speeds of $33\frac{1}{3}$ and 45 r.p.m. made it necessary for the B.B.C. to design a reproducing desk specifically for these recordings. This desk incorporates a number of special features. In the B.B.C. it is comparatively rare for a fine-groove recording to be reproduced in its entirety, but quite common for short excerpts to be used, It must therefore be possible to find a passage quickly and to start reproduction on a chosen word or note. Some form of groove-locating device and a quick-starting device* are therefore required and these have been provided in the design of this equipment.[40]

An optically projected scale has been provided for quick groove location, but with long-playing records this is not in itself sufficiently accurate since it is necessary to locate a precise point within one turn of a groove. The final positioning of the pick-up is determined by rotating the disc forwards or backwards, with the pick-up resting on it, and listening to the output; the turntable is then retracted, and it and the disc are raised to make contact with the pick-up at the instant when the reproduction is required.

The B.B.C. is probably unique among large broadcasting organizations in that less than 50% of its programme output is recorded. Nevertheless during 1960 recordings were made on 79 000 discs and some 108 000 reels of magnetic tape. The standard reel holds 2 400 ft of tape. The majority of th[e] recordings were made for the Transcription Service, whic[h] [dis]tributes recordings of B.B.C. programmes to oversea b[road]casting organizations.

One of the advantages of magnetic recording is, of co[urse,] that the tape can be used again once the original record[ing is] no longer required. The B.B.C. has therefore established a [tape] reclamation unit for testing and repairing tapes; it recond[itions] and returns to service some 1 100 reels of tape each week.

* British Patent No. 815230.
British Patent No. 702090(South African Broadcasting Corporation).

(14) DOMESTIC RECEIVERS

The most important change in receiver design in the p[eriod] under review has been the introduction by the indust[ry of] receivers for the new v.h.f. f.m. sound broadcasting system. [The] high rate of purchase tax has made the evolution of econo[mical] designs more than ever necessary, and such refinements as [push-]button and motor-assisted tuning have virtually disappeared. [In] a great many sets the a.c./d.c. technique has been used to sa[ve the] cost of a mains transformer, and valves have been design[ed to] operate efficiently at the reduced h.t. voltage which this nece[ssarily] entails and with their heaters connected in series. The al[l-glass] miniature type of valve is widely used both in mains-op[erated] and battery receivers.

Most post-war receivers have good sensitivity and reaso[nably] good selectivity, but their standard of audio reproducti[on is] limited by the need for sharpening the response of the [i.f.] circuits to give sufficiently good selectivity for use in th[e con]gested medium-wave band. This need does not a[rise in] v.h.f. f.m. receivers, and a very marked improvement in [audio] quality has been achieved in many of them. The present [trend] is to produce receivers with facilities for v.h.f. recepti[on in] Band II as well as for reception on long wavelengths, m[edium] wavelengths and, in some cases, short wavelengths. Co[mbined] television and v.h.f. receivers are also becoming common, [and] there are some receivers for v.h.f. reception only. Progre[ss has] been made in the standardization of intermediate frequenci[es. An] intermediate frequency of 10·7 Mc/s is widely used in v.h.[f.] receivers; medium- and long-wave sets commonly use an i[nter]mediate frequency of around 470 kc/s, this value being di[ctated] partly by interference considerations and partly by the n[eed to] move self-generated whistles, if they cannot economica[lly be] eliminated, to the less frequently used parts of the [tuning] range.

Some progress has been made in reducing the numb[er of] different types of valves used in domestic receivers, alt[hough] standardization has not been achieved.

The popularity of portable receivers has been one of th[e out]standing features of recent years. Many types were pro[duced] which could be used either with batteries or with an a.c. o[r d.c.] mains supply; portables for use with batteries only were a[lmost] invariably of the 'all-dry' type, in which the l.t. accumulat[or is] superseded by a dry battery.

These types of receiver are now being superseded in thei[r turn] by models in which the valves have been replaced by trans[istors.] By the use of miniature components, printed-circuit wirin[g and] small batteries, the size and weight of these models have [been] greatly reduced, and many are small enough to be slippe[d into] the pocket. Most of these receivers have their own in[ternal] ferrite-rod aerial, but provision is sometimes made for th[e use] of an external aerial in areas where reception is difficult be[cause] of low field strength or other causes.

The use of transistors has now spread to table-model rec[eivers] because the current consumption is so small that the compli[cation] of providing for operation from the mains supply can b[e dis]pensed with. These receivers, although usually of norm[al]

able models in order to provide good-quality reproduction, transportable in that they can be carried from room to n without the need for a mains-supply socket being available. this reason they are sometimes described as 'cordless'.

number of these transistorized table models now include 'ision for reception of the v.h.f. sound service; a welcome 'ency is the spread of this development to portable receivers, ough an external aerial is usually necessary for v.h.f. reception ie fringe areas of the transmitting stations.

ortable radiograms and record reproducers are now featured any manufacturers' ranges and are mostly fitted with 4-speed tables and interchangeable pick-up heads suitable for p.m. and long-playing records.

the field of car radio, the most important development has the introduction of receivers and power-supply units using sistors. This has led to considerable economy in power umption and has enabled the expensive and bulky power incorporating a transformer and vibrator to be dispensed . There are no British car radios covering the v.h.f. band ough imported models are available.

(15) INTERNATIONAL CO-OPERATION

ı the field of international relations, the B.B.C. participates in work of a number of organizations, and is represented on ierous committees and study groups. International co-opera- can naturally assist in the solution of many problems in idcasting; for some of them it is indispensable. During the od under review many links with organizations oversea have ı broken by the war and since re-established. The Inter- onal Broadcasting Union, formed by the broadcasting inizations in Europe in 1925, came to an end with the break of war. It has, unfortunately, not proved possible e the war to form a single comprehensive union of all the idcasting organizations in the European region on account political difficulties. In Western Europe, the European adcasting Union founded in 1950 has replaced the I.B.U.; an organization embracing the whole field of sound and vision broadcasting, including programme and legal matters vell as technical ones. The Director General of the B.B.C. elected President of the Union in 1950 and was re-elected hat office in 1955. A member of the Engineering Division of B.B.C. has been Chairman of the Technical Committee since 2. The E.B.U. has its headquarters in Geneva; the Technical tre is in Brussels with a monitoring and measuring station urbise, in the south-west of Belgium. In Eastern Europe e is a separate union known as the International Organi- on of Radio and Television (O.I.R.T.) which has its dquarters in Prague and includes among its members most he countries of Eastern Europe and others in the Far East. re is some exchange of information and co-operation between two Unions on technical matters.

he B.B.C. also participates in the work of the International communication Union (I.T.U.) and in that of its two nanent consultative committees—the International Radio sultative Committee (C.C.I.R.) and the International Tele- ph and Telephone Consultative Committee (C.C.I.T.T.). se organize studies and issue recommendations and informa- on technical and operating problems. Whenever broad- ing interests are involved, the B.B.C. sends representatives he conferences, either as members of a United Kingdom gation or as independent observers.

he B.B.C. is also represented on the U.K. committee of International Special Committee on Radio Interference S.P.R.) organized by the International Electrotechnical nmission (I.E.C.); the latter body is concerned with standards for all electrical equipment, and the former with the control and suppression of electrical interference caused by such equipment.

(16) INTERNATIONAL EXCHANGE OF PROGRAMMES

The number of sound programmes exchanged 'live' between the B.B.C. and broadcasting organizations in other countries greatly increased during the period under review. Three methods are used for such relays:

(*a*) For exchanges with the nearer European countries, international programme lines incorporated in the telephone network operated by the telephone and telegraph administrations.

(*b*) For more distant parts of the world, radiotelephone circuits operated in many cases by the same administrations.

(*c*) Where the required programme is broadcast on short wavelengths in the country of origin it may be picked up directly at a receiving station operated by the broadcasting organization relaying the programme.

In the case of methods (*a*) and (*b*), the technical arrangements for relays to and from this country are made in co-operation with the Post Office. The technical quality and reliability of these relays have benefited considerably from improvements in the lines and other equipment used. For relays from North America, the first transatlantic telephone cable, commissioned in 1956, is now used almost exclusively; by using two of the circuits in the cable, a channel with a nominal bandwidth of 6·4 kc/s can be obtained for the transmission of music and other high-quality programmes. Many programmes are received from broadcasting stations abroad by direct radio reception at Tatsfield, the B.B.C. Receiving and Frequency Measuring Station in Surrey—notably the European-language transmissions of *The Voice of America*, which are received on short wavelengths from the United States and re-broadcast by the B.B.C. on medium and short wavelengths. A fourth means of exchanging programmes between countries, which is being increasingly used, consists of recordings (transcriptions) either on discs or magnetic tape. The B.B.C.'s Transcription Service distributes annually some 60 000 recordings of B.B.C. programmes to oversea broadcasting organizations.

The great expansion in world communications since pre-war days has widened enormously the field from which programmes can be drawn, but it is still necessary to improvise special circuits from the less highly developed areas as, for example, during the Royal tour of the Commonwealth in 1953–54.

The increase in the number of sound programmes exchanged between the United Kingdom and other countries, excluding transcriptions, is shown in Table 2.

Table 2

	1939 total	1960				
		By radio-telephone	By trans-atlantic cable	Via Tatsfield	By line	Total
Incoming	571	1 209	763	12 859	2 975	17 806
Outgoing	886	324	1 314	—	3 394	5 032

(17) CO-OPERATION IN THE BRITISH COMMONWEALTH

There has been a steady expansion of broadcasting in the Commonwealth countries during the period under review and a rapid development of broadcasting in the Colonies, particularly since 1949 when the British Government decided to make funds available for this purpose under the Colonial Development and Welfare Act.

All the Commonwealth countries have well-developed sound broadcasting services, and all except South Africa have television. In New Zealand television transmissions are at present on an experimental basis only. Thirty-two Colonial territories have sound broadcasting services and three so far have television; seven more have the introduction of television under consideration.

In 1945 the first Commonwealth Broadcasting Conference was held in London; it was called in order that the major broadcasting organizations of the United Kingdom, Australia, Canada, New Zealand, South Africa and India should be able to review their co-operation during the war years and to consult with each other how best this co-operation could be continued and developed in time of peace. Parallel with the meetings of the main Conference a technical committee covered the same field on the engineering and scientific side. The success of this Conference led to the holding of another, also in London, in 1952* and a third in Sydney, Australia in 1956.* A further similar Conference was held in New Delhi in January, 1960.* These Conferences have provided a valuable means of exchanging information and have proposed a number of practical steps to ensure still closer co-operation and mutual assistance in all fields of broadcasting.

Since the war, Colonial broadcasting has created a growing need for experienced broadcasters and engineers to help in establishing and running the new services until trained replacements are available. The B.B.C. has assisted in these projects by seconding staff, on request from the Colonial Office, and by accepting suitable staff from the Colonies and from the newer Commonwealth countries for training by the B.B.C. Assistance has also been given in surveying the various territories in the planning stage in order to provide expert advice on the creation and expansion of a broadcasting service.

In 1960 the number of B.B.C. staff on secondment to the Colonial Office had risen to 46, comprising 25 technical and 21 programme and administrative staff.

This rapid and large-scale development has helped to create an important market for British manufacturers of broadcasting equipment of all kinds ranging from complete transmitting stations and studio centres to recording equipment and receivers.

(18) FREQUENCY MEASUREMENT

Facilities for the accurate measurement of the frequencies of radio transmitters are essential to the functioning of international frequency allocation plans. Official centres have been established for frequency-measurement purposes, and a list of these is published by the I.T.U. At present there are 100 stations carrying out frequency measurement in 27 countries.

In the European Zone, these centres are supplemented in the field of broadcasting by international measuring stations of the E.B.U. at Jurbise, in Belgium, and of the O.I.R.T. in Prague. Individual broadcasting organizations in many countries also operate their own centres, such as the B.B.C. Receiving and Frequency Measuring Station at Tatsfield.[43]

The accuracy of measurement has been steadily improved. Since frequency is a function of time and its unit is the cycle per second, the accuracy with which it can be determined depends upon the accuracy of time determination. Much interesting work has recently been done in observatories and laboratories to reduce the error with which the uniform 'flow' of time can be determined. Special Uniform Time signals are now radiated by the Time Department of Greenwich Observatory (and in the United States by the U.S. Naval Observatory) which enable a standard frequency to be related to Uniform Time in comparison periods of a day or more, with a calculated error of not more than ± 3 parts in 10^9.

* Pakistan, Ceylon, Malaya and Ghana were also represented at one or more of these conferences.

At Tatsfield, precision measurements were made in 1… an accuracy of ± 2 parts in 10^7, which had been impro… 1959 to ± 1 part in 10^8. In terms of time this is equiva… a clock which does not gain or lose more than one sec… approximately $3\frac{1}{4}$ years. In favourable circumstances, me… ments can be made to the accuracy of the standard, i.e. $\pm$… in 10^9, after correction by reference to the Greenwich… Service Bulletin.

Between 450 and 500 frequency measurements are… daily at Tatsfield of B.B.C. transmitters alone. Each lon… and medium-wave frequency is checked twice daily and so … v.h.f. sound and television frequencies. Short-wave freq… used in the B.B.C.'s External Services are checked at th… of each transmission and at three-hourly intervals the… Up to 400 measurements a day are also made of foreign … missions on long and medium wavelengths, short wave… and very high frequencies.

(19) RELAY EXCHANGES (WIRE BROADCASTIN…

The development of wire broadcasting by a number of f… the United Kingdom has fulfilled a need for simple and … reception facilities, especially in areas where direct recep… unsatisfactory and in blocks of flats and in hotels where in… aerials are impracticable. This service is thus comple… to the normal broadcast transmissions.

Sound programmes are generally distributed by wire … frequencies, and subscribers are provided only with a louds… and a programme selector switch. In a few cases a carrier … is used, operating between 62 and 140 kc/s, and trans… several programmes over a single pair of wires. This r… special receiving equipment. In some other countrie… grammes are superimposed on the electricity mains or … telephone system, but these methods are not permitted he…

In 1938 there were 325 sound relay exchanges in op… with a total of 256 000 subscribers. At the end of 19… total number of relay exchanges was 552 (207 for soun… 64 for television only and 281 for sound and televisio… total number of subscribers was 1 074 432 including … television subscribers.

(20) ENGINEERING TRAINING

The problem of maintaining a sufficient number of … technical staff became acute early in the war, when te… staff were called to the armed forces in large numbers a… broadcasting services were expanded on a considerable … To deal with this problem, the B.B.C. Engineering T… Department was created. A start was made with so… instructors dispersed among B.B.C. centres througho… country, and from this small beginning grew the central En… ing Training School near Evesham which was opened in 1…

Since the war, the technical requirements of the various … services have continued to expand, particularly in televis… v.h.f. sound broadcasting. This has created a need for add… staff in the engineer and technician grades; inevitably, the… also been losses of fully trained staff to other organizatio… an increasing rate of normal retirements. Engineering … ment and training therefore form an important part … B.B.C.'s activities.

The Engineering Training Department provides train… broadcasting engineering for all types and grades of te… staff.[41] Attention is focused on the application of… principles to the equipment and methods used by the … without attempting to duplicate the basic training in el… engineering given by the universities and technical college…

e training courses are planned and revised as necessary to the requirements of the sound and television services, n vary as technical development proceeds. A feature of raining technique that has aroused considerable interest de the B.B.C. is the presentation of highly technical inforon in such a way that it can be readily understood by staff out advanced technical knowledge or mathematics. Seven-different types of training courses are now given, ranging the basic courses for new recruits in the Probationary nical Assistant and Operator grades to specialized refresher ses for established engineers. Courses, followed by an ination, are also held for staff wishing to qualify for transfer the Technical Assistant to Engineer grades.

n important feature of the Engineering Training School is it is fully residential (with accommodation for 240 students), giving trainees from different parts of the organization the ortunity of meeting and discussing their problems and their . Trainees have also been accepted from oversea broad-ng organizations, mainly those of the Commonwealth tries. Forty-two such trainees were accepted in the year –59 for various periods of training from two to six months. y of these in turn become instructors on their return e.

ne recruitment of fully-trained engineers is difficult in the of intense competition. Considerable effort has therefore devoted to recruiting boys from the sixth forms of schools about 18, who have studied up to G.C.E. Advanced Level athematics and physics. This method is likely in future to luce the bulk of the B.B.C.'s intake of Probationary Tech-l Assistants and Operators.

scheme has been introduced to provide a sandwich course selected 18-year-olds leading to Higher National Diploma then to Graduate Membership of The Institution of Elec-l Engineers. This is additional to the long-established me for the training of graduate apprentices.

(21) ELECTRICAL INTERFERENCE

ne B.B.C. is vitally concerned with all kinds of interference ch may spoil reception of the programmes by its public, the ners and viewers. The enormous growth in the number of rical appliances used in the home, in addition to greatly eased use of industrial and electro-medical apparatus, has e local electrical interference a major hazard to reception, icularly since the war. In some areas, severe interference so experienced from overhead power lines. The B.B.C. is efore intimately concerned with the technical investigation nis problem and works in close collaboration with the Post ce, The Institution of Electrical Engineers, the British Stan-ls Institution, and other interested organizations, including C.I.S.P.R. (International Committee on Radio Interference). is satisfying to be able to record that, as a result of careful nical investigations and consultation, the Postmaster General been able to make Regulations under the Wireless Telegraph 1949, for the control of interference caused by the ignition ems of motor vehicles and stationary installations using rnal combustion engines (1953) and by small electric motors refrigerators (1955). Interference from electro-medical and istrial heating equipment is under consideration by the master General's Advisory Committee.

nfortunately, the fitting of ignition-interference suppressors lder (pre-1953) vehicles has not been made compulsory, and e vehicles, though diminishing in numbers, continue to cause ous interference to television and v.h.f. sound broadcasting, icularly in areas of low signal strength.

ne investigation of complaints of electrical interference is undertaken by the Post Office; the number of complaints investigated has steadily fallen during the past four years as is shown by the following figures:

1956	162 251
1957	135 994
1958	123 331
1959	107 877

This reduction is no doubt due partly to the reduced use of long-wave and medium-wave receivers, now that almost the whole country is served with all three programmes on v.h.f. transmissions. The v.h.f. services are much less susceptible to interference. It is also partly due to a transfer of public interest to television. It seems likely, however, that some part of the welcome reduction in the number of complaints is a result of the introduction of the regulations and of the greater awareness of the problem among manufacturers and users of electrical appliances.

International agreement, and where possible standardization, in matters relating to electrical interference is important. At a Plenary Meeting of the C.I.S.P.R. held in Holland in 1958, agreement was reached on standard performance specifications for interference measuring sets covering the ranges 0·15–30 Mc/s and 25–300 Mc/s. This is an important achievement from the point of view of the electrical industry because it means that, although different limits may be applied in different countries, the measurements are made with the same type of equipment so that the manufacturer can satisfy himself that his product conforms with the regulations (or voluntary specifications) adopted in any country to which he intends to export it. In view of the importance of future developments in Europe in Bands IV and V, a new specification is now to be drawn up for frequencies between 300 and 1 000 Mc/s.

It is not to be expected that international agreement will be reached soon, if ever, on the limits to be applied to the interference produced by electrical appliances of all kinds; this is because the sound and television broadcasting services in different countries find it necessary to protect different values of field strength in accordance with local conditions. Nevertheless, a great deal of useful information has been collected on the subject of limits and also on the difficult problem of obtaining a satisfactory degree of suppression without infringing the safety rules adopted by the C.E.E. (International Committee for Electrical Safety). This information is contained in a report published by the Central Office of the International Electrotechnical Commission.[42]

(22) ACKNOWLEDGMENTS

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In conclusion the B.B.C. wishes to pay tribute to the technical achievements of the radio industry over the last 21 years, which have made a major contribution to the progress described in the review. The B.B.C. also acknowledges the help and co-operation of the Post Office in the development of its plans in the technical as in other fields. Valuable work has been done by the several international organizations mentioned herein and also by the British Standards Institution and other professional bodies in this and other countries. Finally The Institution itself has played an important part in maintaining professional standards, and in encouraging technical developments and making them known to engineers throughout the world.

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(24) APPENDICES

) B.B.C. Long and Medium Wavelength Transmitting Stations

Regional Programmes, 1939

Station	Frequency	Power	Programme
	kc/s	kW	
oorside Edge ..	668	65	Northern
esterglen ..	767	65	Scottish
rghead ..	767	60	Scottish
ashford	804	65	Welsh
nmon	804	8	Welsh
ookmans Park ..	877	60	London
snagarvey ..	977	100	Northern Ireland
oitwich	1 013	65	Midland
art Point ..	1 050	100	West
agshaw	1 122	80	North
edmoss	1 285	3	Scottish
evedon	1 474	20	West

e coverage of the above stations was 89% of the population of the United dom. (Total population, 1931 census, 46 180 000.)

Home Services, 1959

Station	Frequency	Power	Programme
	kc/s	kW	
oorside Edge ..	692	150	Northern
hitehaven ..	692	2	Northern
romer	692	2	Northern
urghead	809	100	Scottish
edmoss	809	5	Scottish
esterglen ..	809	100	Scottish
umfries	809	2	Scottish
enmon	881	8	Welsh
owyn	881	5	Welsh
ashford	881	100	Welsh
rexham	881	2	Welsh
rookmans Park ..	908	140	London
art Point ..	1 052	120	West
arnstaple ..	1 052	2	West
roitwich ..	1 088	150	Midland
ostwick	1 088	7·5	Midland
isnagarvey ..	1 151	100	N. Ireland or Northern
ondonderry ..	1 151	0·25	
carborough ..	1 151	2	
agshaw	1 151	100	
artley	1 457	10	West
righton	1 457	2	West
levedon	1 457	20	West
olkestone ..	1 457	1	West
exhill	1 457	2	West
edruth	1 457	2	West
arrow	1 484	2	North
amsgate	1 484	2	London

e coverage of the above stations is 93% of the population of the United Kingdom. al population, 1951 census, 50 369 000.)

National Programme, 1939

Station	Frequency	Power
	kc/s	kW
Droitwich	200	180
Brookmans Park ..	1 149	30
Moorside Edge ..	1 149	40
Westerglen	1 149	45

The coverage of the above stations was 93% of the population of the United Kingdom. (Total population, 1931 census 46 180 000.)

Light Programme, 1959

Station	Frequency	Power
	kc/s	kW
Main Transmission:		
Droitwich	200	400
Auxiliary Service:		
Brookmans Park ..	1 214	50
Burghead	1 214	20
Lisnagarvey	1214	10
Londonderry ..	1 214	0·25
Moorside Edge ..	1 214	50
Newcastle	1 214	2
Plymouth	1 214	0·25
Redmoss	1 214	2
Redruth	1 214	2
Westerglen	1 214	50

The coverage of the above stations is 99% of the population of the United Kingdom

Third Programme and Network Three

Station	Frequency	Power
	kc/s	kW
Daventry	647	150
Edinburgh	647	2
Glasgow		
Newcastle-on-Tyne		
Redmoss		
Belfast	1 546	Between 0·25 and 1
Bournemouth ..		
Brighton		
Dundee		
Exeter		
Fareham		
Leeds		
Liverpool		
Plymouth		
Preston		
Redruth		
Stockton-on-Tees ..		
Swansea		

The coverage of the above stations is 69% of the population of the United Kingdom. (Total population, 1951 census, 50 369 000.)

The above figures refer to night-time coverage in the absence of foreign interference; the effective coverage is severely reduced during periods of maximum foreign interference, particularly in the case of the Home Service stations.

(24.2) **B.B.C. Stations Transmitting the Home, Light, Third and Network Three Programmes on V.H.F.**

Station	Frequencies			Effective radiated power of each transmitter	Opening date
	Light	Third/ Network 3	Home		
	Mc/s	Mc/s	Mc/s	kW	
Blaen-plwyf	88·7	90·9	93·1	60	14.10.56
Divis	90·1	92·3	94·5	60	18.3.56
Douglas (I.O.M.)	88·4	90·6	92·8	3·3	9.3.58
Holme Moss	89·3	91·5	93·7	120	10.12.56
Kirk o'Shotts	89·9	92·1	94·3	120	30.11.57
Llanddona	89·6	91·8	94·0	6	20.12.58
Llangollen	88·9	91·1	93·3	7	20.12.58
Meldrum	88·7	90·9	93·1	60	29.3.56
North Hessary Tor	88·1	90·3	92·5	60	7.8.56
Orkney	89·3	91·5	93·7	25 max*	22.12.58
Peterborough	90·1	92·3	94·5	22 max*	5.10.59
Pontop Pike	88·5	90·7	92·9	60	20.12.55
Rosemarkie	89·6	91·8	94·0	6	12.10.58
Rowridge	88·5	90·7	92·9	60	4.6.57
Sandale (Carlisle)	88·1	90·3	94·7‡ 92·5†	120	18.8.58
Sutton Coldfield	88·3	90·5	92·7	120	30.4.57
Tacolneston (Norwich)	89·7	91·9	94·1	120	30.4.57
Thrumster	90·1	92·3	94·5	10 max*	1.3.60
Wenvoe	89·95	96·8	94·3‖ 92·125§	120	27.5.57
Wrotham	89·1	91·3	93·5	120	2.5.55

* Directional aerial.
† Scottish Home Service.
‡ North Home Service.
§ West of England Home Service
‖ Welsh Home Service.

The population coverage of the above stations is 97·3% of the population of the United Kingdom (250 μV/m contour). (Total population, 1951 census, 50 369 000.)

(24.3) **Proposed B.B.C. V.H.F. Satellite Transmitting Stations for the Home, Light and Third Programmes**

Stage I

Fort William
Galashiels area
Berwick-on-Tweed
Llandrindod Wells area
Kinlochleven
Oban
Oxford/Berkshire (Four sound programmes)
Redruth
Les Platons, Channel Islands
Londonderry

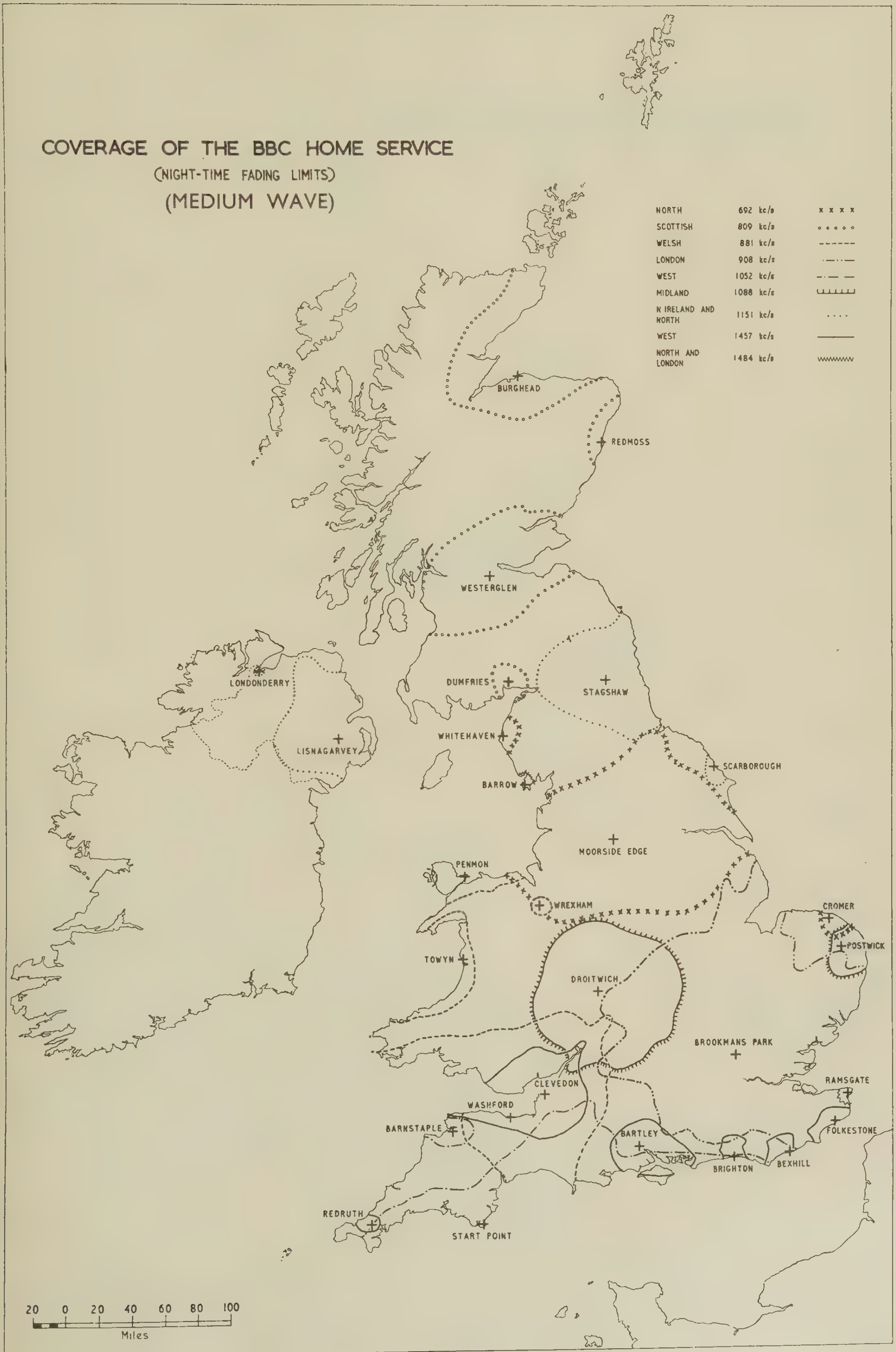
Stage II

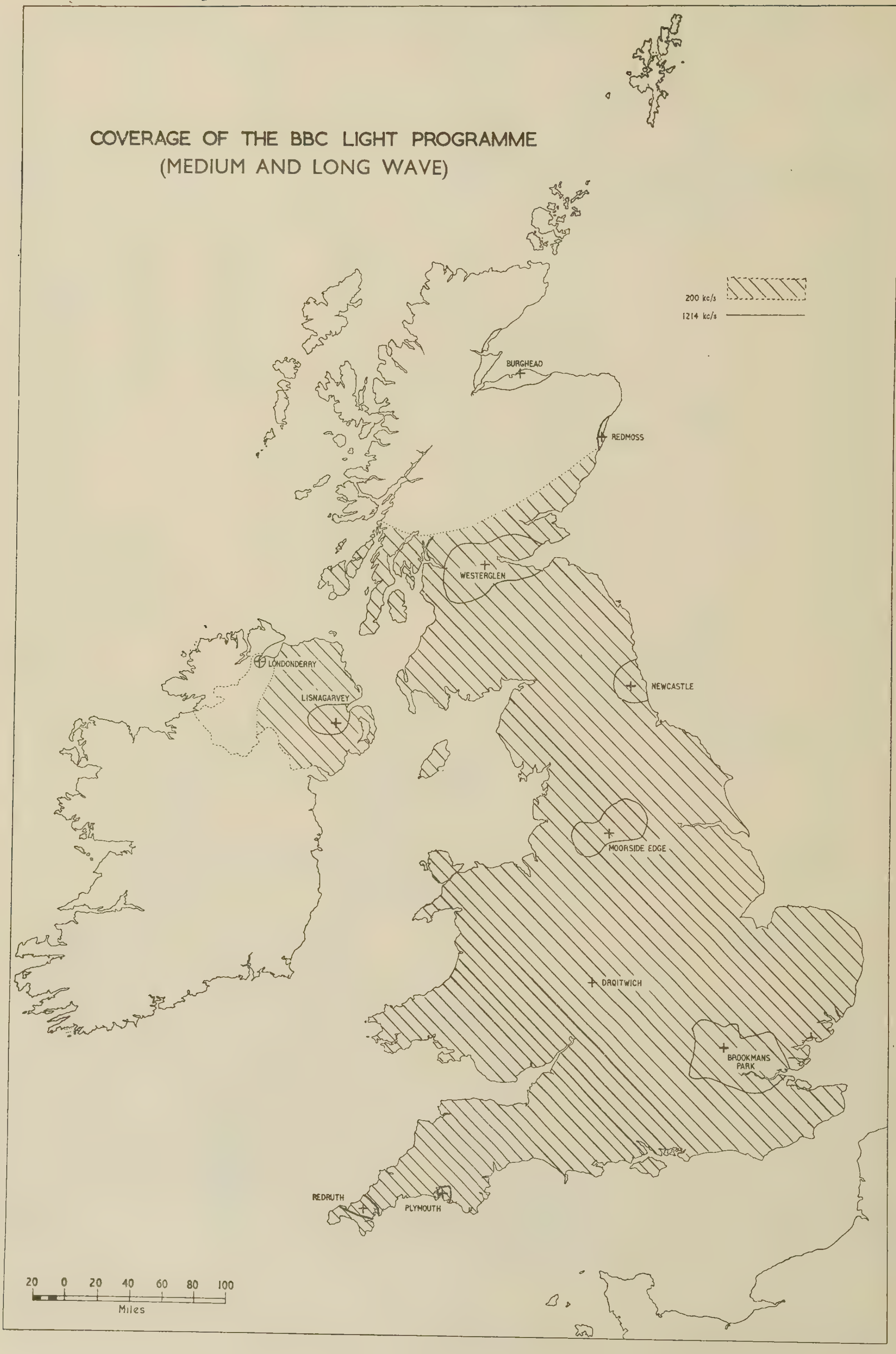
Forfar, Angus
Grantown-on-Spey
Lewis
Pitlochry/Aberfeldy
Shetland
Skye
East Lincolnshire
Enniskillen
Pembroke/Milford Haven
Sheffield
South-West Scotland

Until sites have been chosen and technical operating [con]ditions agreed, it is not possible to predict exactly the imp[rove]ments in coverage which the above stations will achieve. [It is] estimated, however, that the 21 v.h.f. stations will increas[e] coverage by some 990 000 (1·96%), and give improved se[rvice] for a further 1 200 000 people.

Stage 1 is due to be completed during 1961 and 1962; [con]struction of the stations in Stage II is proceeding concur[rently] and it is expected that most of the stations will be complet[ed by] the end of 1963 and the remainder in the spring of 1964.

(24.4) Coverage of the B.B.C. Services

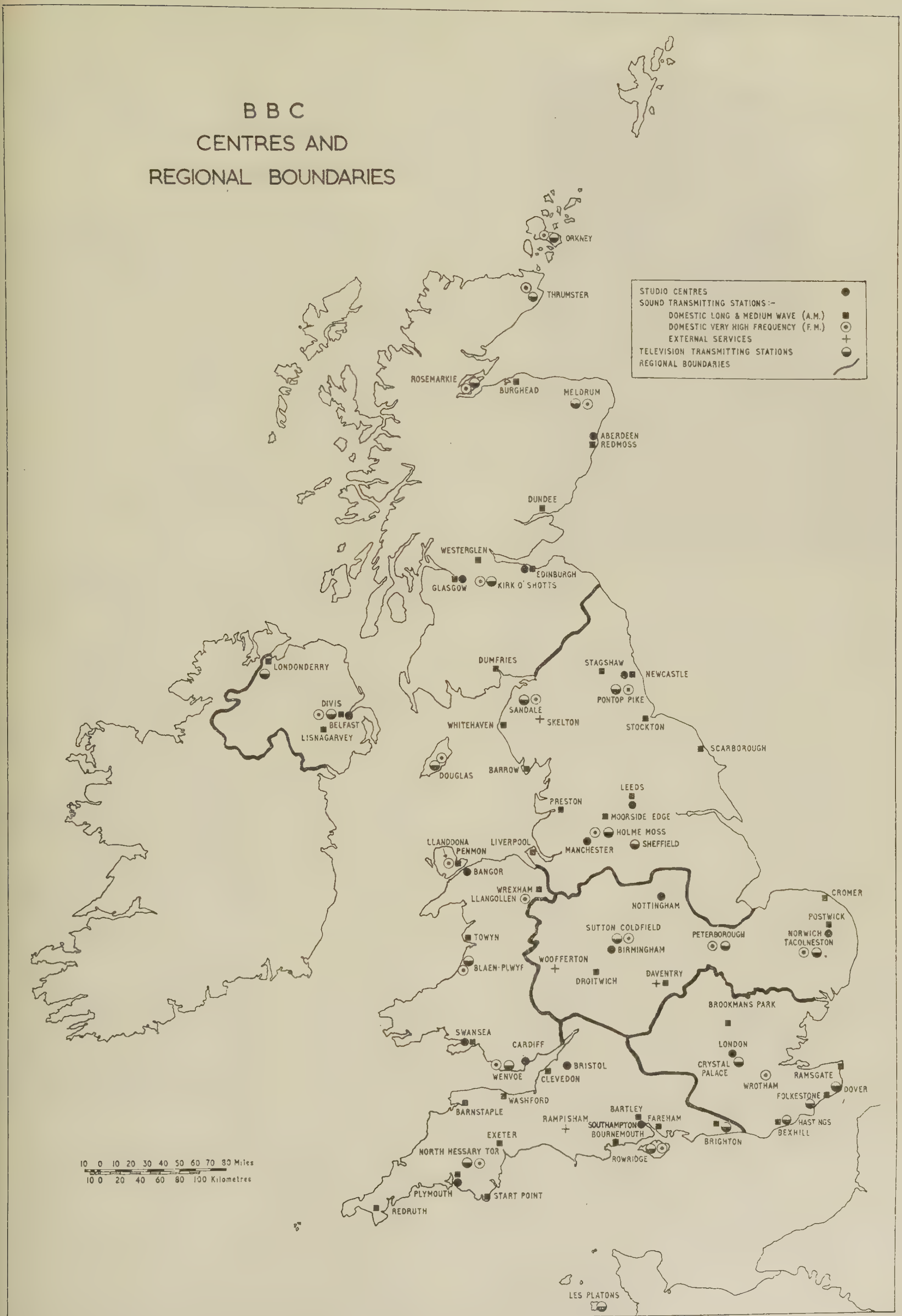




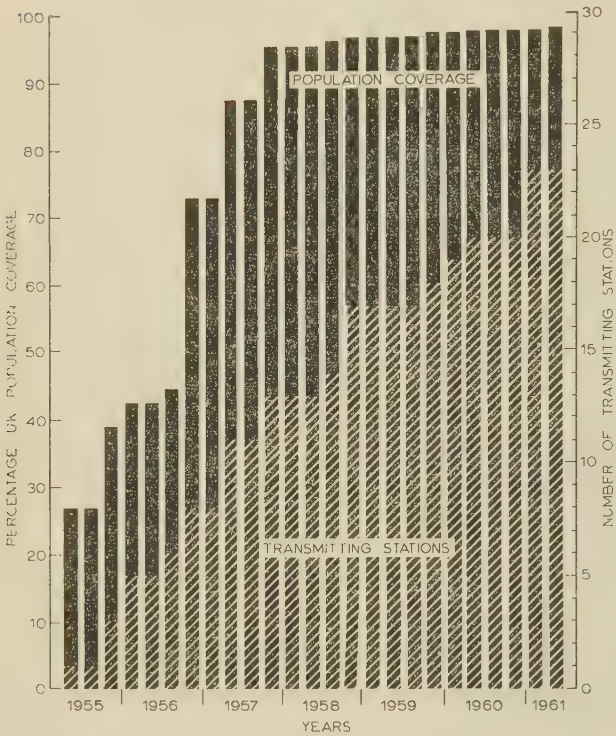
COVERAGE OF THE BBC LIGHT PROGRAMME
(MEDIUM AND LONG WAVE)


COVERAGE OF THE BBC THIRD PROGRAMME
AND NETWORK THREE
(MEDIUM WAVE)
647 kc/s
(NIGHT-TIME FADING LIMITS)
1546 kc/s
REDMOSS
DUNDEE
EDINBURGH
GLASGOW
NEWCASTLE
STOCKTON
BELFAST
PRESTON
LEEDS
LIVERPOOL
DAVENTRY
SWANSEA
FAREHAM
EXETER
BOURNEMOUTH
BRIGHTON
PLYMOUTH
REDRUTH
20 0 20 40 60 80 100
Miles

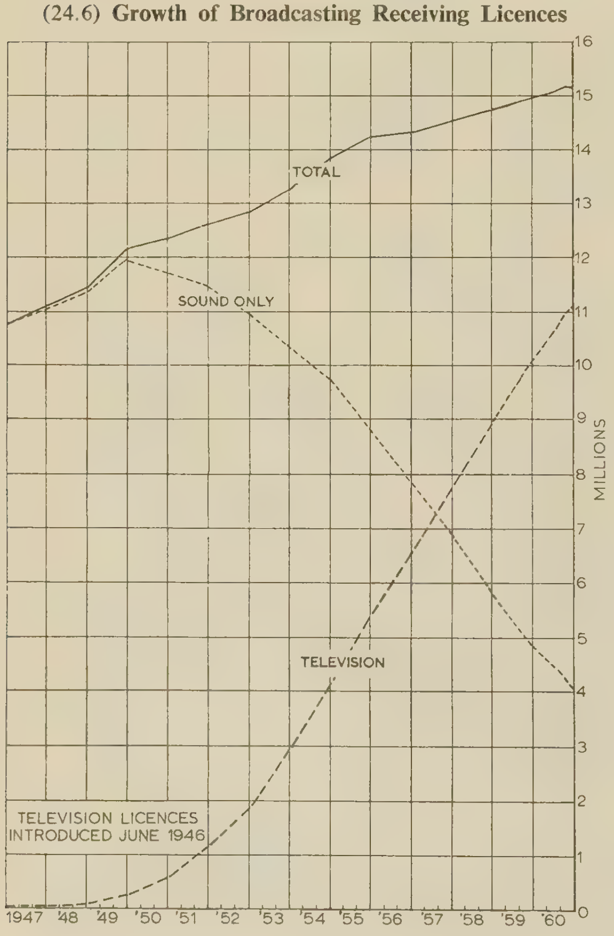

EXISTING STATIONS
LOW-POWER SATELLITE STATIONS
PLANNED OR UNDER CONSTRUCTION
STAGE I +
STAGE II ×
AREAS NOT YET COVERED
COVERAGE OF BBC VHF SOUND BROADCASTING
SHETLAND
ORKNEY
THRUMSTER
LEWIS
ROSEMARKIE
SKYE
MELDRUM
GRANTOWN-ON-SPEY
FORT WILLIAM
KINLOCHLEVEN
FORFAR
PITLOCHRY/ABERFELDY
OBAN
KIRK O' SHOTTS
BERWICK-ON-TWEED
GALASHIELS AREA
LONDONDERRY
DIVIS
SOUTH-WEST SCOTLAND
PONTOP PIKE
SANDALE
ENNISKILLEN
DOUGLAS
HOLME MOSS
SHEFFIELD
LLANDDONA
EAST LINCOLNSHIRE
LLANGOLLEN
PETERBOROUGH
SUTTON COLDFIELD
TACOLNESTON
BLAEN-PLWYF
LLANDRINDOD WELLS
OXFORD/BERKS.
PEMBROKE/MILFORD HAVEN
WENVOE
WROTHAM
DOVER
NORTH HESSARY TOR
ROWRIDGE
REDRUTH
20 0 20 40 60 80 100
Miles
LES PLATONS


B B C
CENTRES AND
REGIONAL BOUNDARIES
STUDIO CENTRES
SOUND TRANSMITTING STATIONS :-
DOMESTIC LONG & MEDIUM WAVE (A.M.)
DOMESTIC VERY HIGH FREQUENCY (F.M.)
EXTERNAL SERVICES
TELEVISION TRANSMITTING STATIONS
REGIONAL BOUNDARIES
10 0 10 20 30 40 50 60 70 80 Miles
10 0 20 40 60 80 100 Kilometres
ORKNEY
THRUMSTER
ROSEMARKIE
BURGHEAD
MELDRUM
ABERDEEN
REDMOSS
DUNDEE
WESTERGLEN
EDINBURGH
GLASGOW
KIRK O' SHOTTS
LONDONDERRY
DIVIS
BELFAST
LISNAGARVEY
DUMFRIES
STAGSHAW
NEWCASTLE
PONTOP PIKE
SANDALE
SKELTON
WHITEHAVEN
STOCKTON
SCARBOROUGH
DOUGLAS
BARROW
LEEDS
PRESTON
MOORSIDE EDGE
HOLME MOSS
SHEFFIELD
LLANDDONA
PENMON
LIVERPOOL
MANCHESTER
BANGOR
WREXHAM
LLANGOLLEN
NOTTINGHAM
CROMER
POSTWICK
SUTTON COLDFIELD
PETERBOROUGH
NORWICH
TACOLNESTON
TOWYN
BIRMINGHAM
BLAEN-PLWYF
WOOFFERTON
DROITWICH
DAVENTRY
BROOKMANS PARK
SWANSEA
CARDIFF
LONDON
CRYSTAL PALACE
WENVOE
CLEVEDON
BRISTOL
WROTHAM
RAMSGATE
DOVER
FOLKESTONE
WASHFORD
BARNSTAPLE
BARTLEY
RAMPISHAM
SOUTHAMPTON
FAREHAM
BOURNEMOUTH
HASTINGS
BEXHILL
BRIGHTON
EXETER
NORTH HESSARY TOR
ROWRIDGE
PLYMOUTH
START POINT
REDRUTH
LES PLATONS

(24.5) **V.H.F. Population Coverage**



(24.6) **Growth of Broadcasting Receiving Licences**



(24.7) **B.B.C. Sound-Recording Facilities**

		Film (Philips–Miller)	Steel tape (Marconi–Stille)	Direct disc	P
1939	Static	2	6	6	
	Mobile	—	—	4	-
1959	Static	—	—	68 (42)	24
	Mobile (in vehicles)	—	—	29	8
	Midget (portable)	—	—	—	22

Figures in brackets refer to 1960.

A further 270 magnetic tape recorders are used for rehearsals.

(24.8) **Some Outstanding Dates in the Development of Sound Broadcasting Since 1939 (Omitting the War Yea**

1st September, 1939 — Single programme (Home Service) re pre-war National and Regional Progra

29th July, 1945 — Post-war programmes introduced—Hor vice and Light Programme.

29th September, 1946 — Third Programme introduced.

2nd May, 1955 — V.H.F. sound broadcasting service (first transmitting station, Wrotham,

1st September, 1957 — Network Three introduced (using the Programme transmitters outside Programme hours).

13th and 14th January, 1958 — First modern stereophonic test transm (London transmitters only; 11th an May, 1958 from transmitters thro the United Kingdom).

18th October, 1958 — Regular fortnightly experimental stereo transmissions began (using Network and television sound transmitters th out the United Kingdom).

24th June, 1959 — B.B.C. satellite transmitting station announced (Stage I); includes 1 sound stations.

20th May, 1960 — Stage II of B.B.C. satellite trans stations plan announced; 11 static v.h.f. sound included.

8th September, 1960 — Announcement by Postmaster Gene Committee of Inquiry into the of Sound and Television Broad (Pilkington Committee).

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TELEVISION BAND COMPRESSION BY CONTOUR INTERPOLATION

By Professor D. GABOR, Dr.-Ing., F.R.S., F.Inst.P., Member, and P. C. J. HILL, Ph.D.

(*The paper was first received* 17*th June, and in revised form* 28*th November*, 1960.)

SUMMARY

ıd saving in television transmissions can be achieved by utilizing dancies in single lines, between lines, between fields and between s. The method of contour interpolation exploits the last two. ɔased on the facts that (*a*) field and frame frequencies in conven- television transmission had to be chosen with a view to reducing r rather than for conveying more information, (*b*) the eye fixes y on contours which are usually the edges of objects, (*c*) these urs are usually smooth enough to allow interpolation over ine spacings, and (*d*) changes from one picture to the next come t mostly by the horizontal motion of objects which are sufficiently rm to allow interpolation over at least two frame intervals. is almost no loss in information or picture quality if the inter- frame is suppressed in the transmission and reconstructed in eceiver by interpolation between the lines of the transmitted and there is not much loss if only one field in four is transmitted. e method of contour interpolation the reconstructed edges are arp as the originals and appear in their correct positions, i.e. in ɔositions which they would occupy if the edges were straight in sections, and if their motion were uniform for short times. waveband gain can be estimated as 4 : 1 without appreciable ioration in picture quality, and 8 : 1 if some deterioration is ved in the case of rapid and vertical motions. In combination other methods which utilize redundancies in single lines and h by themselves achieve a gain of 3 : 1, total compression ratios : 1 to 24 : 1 appear feasible.

e principle was realized and tested in a photo-mechanical model ing at low speeds. Electronic realizations are proposed and ssed.

1) INTRODUCTION. MAXIMUM REDUNDANCY IN TELEVISION TRANSMISSION

has been obvious for a long time that the standard system elevision is a very wasteful method of transmitting visual rvation. It will be of interest to start by estimating the imum saving in waveband which could conceivably be zed if we were not bound by any economic considerations ie terminal equipment.

ie bandwidth in television has been chosen with a view to ɔst satisfying the foveal, i.e. the maximum acuity of vision. 400-line 21 in picture, viewed from a distance of 5 ft a line ends about 2′ of arc, which is about twice the foveal resolu- In practice, with a 2 : 1 interlace the vertical definition esponds to only about $0{\cdot}6 \times 400 = 240$ lines (with a Kell ɔr of 60%), which brings the effective line width to about times the eye resolution limit. Even this rather unsatis- ɔry compromise is achieved at the cost of an enormous nel capacity. If we want to estimate the maximum saving h might be effected, we must first remember that the fovea ends an angle of only 1°, which at a distance of 5 ft corre- ds to rather less than 1/200 of the screen area.

we set out to make use of the large potential saving, which is ested by the smallness of the fovea, we must first realize a facts about human vision. Psychologists assure us[1] that eye can accept information at the rate of almost 50 bits/sec. Yet we can 'take in' a whole unknown landscape at a single glance, and we have the subjective impression that we see everything within a solid angle of almost 2π with perfect acuity. The reason is evidently that even the peripheral acuity is sufficient to recognize familiar objects, such as trees, houses, etc., and until the fovea has had time to scan these, we substitute standard trees, etc., for them, out of our store of visual memory, which approximately fill their contours. We become aware of the slowness of our intake only when we are faced with unfamiliar objects, such as complicated anatomical preparates, or a detailed drawing of a jet engine. In such cases we have to scan the picture slowly and rather painfully, line by line, and do not get the feeling that we have really 'taken it in' for several seconds or minutes.

No television system is conceivable which takes full advantage of this limitation of vision, unless by some signals transmitted to the eye muscles it could force its millions of viewers to scan the picture in perfect synchronism. But, in principle at least, we could do almost as well, so long as we are satisfied with transmitting scenes only, or at the most only small unfamiliar objects. The transmitter will then have to be a recognizer of sufficient intelligence to pick out areas containing foliage, grass, crowds, wallpaper patterns and the like, sending to the receiver the approximate contours and code numbers for substituting the nearest standard pattern. The substituted background will be, of course, spurious in detail but sharp, and the fovea will be perfectly satisfied when it scans over it. It is interesting to note that many of these patterns, like foliage, grass, a choppy sea, etc., are just those which by their 'noise-like' character tax ordinary television transmission to the limit of its capacity, and yet they are of very little importance to the information transmitted in the sense of 'telling a story'. After having covered most of the picture with standard substitutes only a small fraction of the area must be left, such as the faces of the principal actors, which must be transmitted in true and full detail. As a rough estimate, in something like 80% of television transmissions this area could be restricted to perhaps 10 foveas, i.e. 5% of the total. But here again, by means to be discussed later, a saving of about 4 : 1 could be effected by making use of the point-to-point redundancy. Making an equal allowance for the background, the information transmitted per picture might thus be reduced to $2{\cdot}5$%, giving a saving of 40 : 1.

A further saving (again at the expense of a minority of scenes) could be effected by transmitting changes only, i.e. holding or repeating the unchanged part of the picture by some storage system in the receiver. This has often been proposed, and the familiar objection that a panning camera would upset the system could be met in principle by embodying in the receiver a system for effecting geometrical transformations, so that only the newly appearing picture parts need be transmitted. There is, however, an absolute limit to this, because the viewers would certainly object if it took more than about 1/4 sec to build up a new picture, instead of 1/25 sec as at present. This gives a potential saving of $6{\cdot}25$: 1 in picture sequences, and hence an overall bandwidth saving of 250 : 1.

We want to make it clear that we do not believe this to be the television system of the future. In all probability it will

tten contributions on papers published without being read at meetings are l for consideration with a view to publication.

f. Gabor is Professor of Applied Electron Physics, and Dr. Hill was formerly Department of Electrical Engineering, Imperial College of Science and Tech- ", University of London. Dr. Hill is with the B.B.C. Research Department.

never be economically justifiable. It is a minor objection that we are still very far from being able to construct a machine with enough visual intelligence to divide up a picture into familiar types of objects such as trees, grass, human figures, etc. There is every hope that this difficult problem, which is of the greatest intrinsic scientific interest, will be solved some time, and one could well afford a very complicated machine at the transmitter end. Picking out the objects, such as the faces, hands, etc., of the principal actors which are of interest for the story appears to call for even higher intelligence, but this could be short-circuited by some special paint, visible to the camera but not to the viewers. The bottleneck is not in the transmitter but in the receiver, which would have to be of such complication that it could never be seriously contemplated as a home receiver. Our object in discussing this ideal picture transmission system in some detail is rather to point out that the estimates of potential compression based on contrasting the 50 bits/sec of visual intake with the 3–5 $\times$ 10^7 bits/sec of television channels are widely unrealistic. This suggests a potential compression of the order of a million, while our rather optimistic estimate is about 250—and this at the cost of excluding a rather important class of transmissions, such as ballet pictures, large orchestras and scientific demonstrations of very unfamiliar objects. 25 : 1 is not even sufficient for transmitting television through telephone lines, which is often mentioned as a practical aim.

(2) PREVIOUS ATTEMPTS AT TELEVISION BANDWIDTH REDUCTION

Shannon's information theory can classify all attempts at bandwidth reduction in communication problems, although in a way which is not completely satisfactory when applied to television systems and other problems in which the receiver is the human sensory system. The situation visualized by Shannon is as follows: A source transmits a range of signs numbered $1 \, . \, . \, i \, . \, . \, N$ with probabilities p_i. If these probabilities were equal, i.e. $p_i = 1/N$, they would require $\log_2 N$ bits/sign for their transmission, assuming a noiseless channel. If the probabilities are unequal, they require on average

$$H = -\sum_{1}^{N} p_i \log_2 p_i \text{ bits/sign}$$

which is always less than $\log_2 N$. A coding method can be found which transmits messages from this source in long sequences with H bits/sign, with an error which can be made arbitrarily small. The bandwidth required to transmit these signs at a certain rate can then be calculated with the well-known Shannon–Tuller formula.

The difficulty in applying Shannon's formula to television transmission lies in the fact that there is no simple answer to the question of what the human receiver will accept as distinguishable communication signs. In the previous Section we have tried to make the tolerance as wide as possible, by assuming that outside the narrow foveal field the eye will accept, for instance, standard grass for the infinite variety of real grass, so that the major part of the picture is defined by outlines, filled with say a few hundred or thousand standard patterns, Such a radical step in reducing the code has not yet been attempted. All attempts of which we know identify the signs i with small groups of adjacent picture points in one field. Most of them consider picture points in one scanning line.

In order to reduce the television problem to Shannon's discrete coding some sort of quantization is required. Goodall[2] has investigated the number of levels acceptable to the eye. It appears that 32 levels are the minimum for gently shaded areas, while much cruder quantization can be allowed at contours and in high-frequency regions. This has suggested a soun band saving to several authors.[3–5] Kretzmer[6] has show the probability distribution of the 32 levels, the so-calle order statistics', is so nearly even that optimum coding only an insignificant gain on the 5 bits/point required by form distribution. For larger gains at least two conse points must be considered, making use of the so-called 'se order statistics'. If one knows the amplitude at a po described by a sign i, the amplitude j at the point $n + 1$ rather narrow conditional probability distribution $p_i(j)$, sharp maximum at $j = i$, i.e. at 'no change'. Schreibne Stoddard[8] have measured the second-order probabilit tribution, and Stoddard obtained a conditional entro 1·24 bits/point for a typical 32-level television picture. is to say that, by making use of the correlation betwe secutive picture points, it is possible to recode the amplit a 4 : 1 reduced waveband.

The suggestions made by Schröter[9] and by Cher Gouriet[10] can be considered as realizations of the gain in the use of point-to-point conditional or 'transition' pro ties. These authors propose scanning the lines at va velocity, fast where there is little change and slowly whe amplitude is steeply changing. Gouriet[11] proposed to m the 'picture detail' by the square of the rate of amplitude and to make the scanning speed inversely proportional t One of the authors (in the Appendix to Reference 1 pointed out that for a very plausible law of transition prob the Cherry–Gouriet method is equivalent to ideal codin point-to-point basis, i.e. it realizes complete equalization rate of information transmission. In practice, this can fully realized, because it is not possible to scan uniform r at infinite speed; 3 : 1 appears to be the largest variati speed which can be realized without instabilities in the c loop system. This corresponds very nearly to a compr ratio of 3 : 1. A further improvement has been sugges Cherry, Prasada and Holloway[12] in an 'open-loop' syst which use is made of the smaller number of quantized l required in regions of high picture detail, and in whic information is sent out again at a practically uniform We will refer to these schemes as 'rate-equalization' me It may be noted that no great improvement is to be ex in these beyond the ratio of 4 : 1, because Schreiber[1 measured 'third-order statistics' and found that the incre gain is disappointingly low. Gouriet's estimate[14] of a po gain of 7 : 1 is rather out of harmony with these estimates. possible that even 4 : 1 is somewhat optimistic, as Stod figure was based on the statistics over a whole picture, a practical difficulties in attempting more than line-by-line c are formidable. Even using line-by-line coding it is very d to make corresponding points exactly coincide after the dec It appears that Youngblood's[15] interesting suggestion of tizing the rate of change instead of the amplitude levels because of this difficulty. Though abrupt level changes line were eliminated, they appeared between one line a next.

The importance of exactly transmitting the positions of amplitude changes, such as occur at edges, at the expense conspicuous detail was recognized at an early stage.[3–] particular interest is the system of 'synthetic highs' pr by Schreibner, Knapp and Kay,[16] in which the positio height of edges is transmitted in a coded form in a rate-equa system, and a low-definition directly-transmitted picture is posed on the sharp contours in the receiver.

While the systems so far described are based on proc the information contained in one horizontal line, some att have also been made to exploit vertical correlations, fro

ne. A scheme in which the next line is predicted and only surprise' difference is transmitted was investigated (together other applications of linear prediction in picture transion) by Harrison.[17] Work on this was continued by Julesz[18] Cunningham,[19] who proposes interpolation in 3 × 3 and 5 sample blocks.

e wish to exploit correlations between consecutive frames is ld as television itself; it would be difficult to name the or who first proposed to leave the static parts of the picture ling and to write in only the changes. The modern lopment of storage tubes, in particular the invention of tive erasure,[20] has made such schemes technically but not t economically possible.

(3) PRINCIPLE OF CONTOUR INTERPOLATION

ır proposals for waveband reduction are based entirely on exploitation of the correlations between consecutive fields consecutive frames.* These are independent of the correla- in the contents of single lines, and hence whatever gain hievable by our system can be multiplied by a factor of f it is combined with a line-by-line compression method.

ght from the start television had to make a concession to flicker sensitivity of the human eye by introducing the interlace, in which 50–60 fields (half-pictures or halfes) are presented instead of 25–30 full frames. By this st one-half of the vertical definition was sacrificed, because in a very carefully adjusted receiver the subjective impression picture with 400 lines composed of two interlaced fields 200 lines is equivalent to not more than 240 lines, and in t receivers to only 200. Thus it can be said that the inter- field contributes almost nothing to the information; half waveband is sacrificed for the suppression of the flicker.

follows that it must be possible in principle to halve the eband without loss of information if an alternative way is d to reduce the flicker. In cinematography 24 full pictures projected, each shown twice, and 16 mm projectors achieve cceptable effect with 16 pictures, each shown three times. etition in television could be achieved by attaching to the iver tube a memory device storing a full picture, or, better, sing a storage tube for the display in which the old picture is ed just before the new one is recorded. A somewhat better lt could be obtained by merging one frame continuously into next, by superposing the new and the old frame in conously changing proportions, as is done for instance in the hau projector. We will refer to this method as 'linear polation'.

l these methods are far too complicated and expensive to mbodied in domestic receivers. At present, one can conplate any sort of complicated information processing only in stations, which ultimately transmit standard interlaced s to the home receivers. It was with this application in that one of the authors (D. G.)[21] proposed transmitting one field out of two, and constructing the missing inter- field by a method called 'contour interpolation', which is rior both to repetition and 'linear interpolation'.

e method is based on the physiological fact that the eye with preference on outlines, i.e. loci of abrupt amplitude ges, and on the common observation that most outlines ontinuous. The principle is explained in Fig. 1.

e two diagrams show, in their first and third lines, the sity distributions along two consecutive lines of a field h has been transmitted and received, while the middle line, h belongs to the interlaced field which was not transmitted,

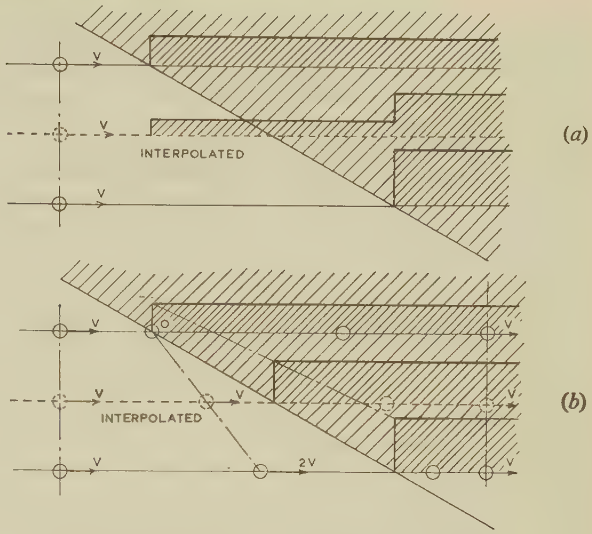


Fig. 1.—Linear and contour interpolation.

(a) Linear interpolation.
(b) Principle of contour interpolation. Two scanning spots are moving on two next-but-one lines, normally with constant speed v. When one of these arrives at an edge it stops and the other moves on with double speed until it has also reached the edge. The intensity in the interpolated line is the arithmetical mean weighted with the scanning speeds.

must be constructed in the receiver by some method of interpolation. Both examples relate to a slanting contour which separates two areas—one dark and the other bright. It may be assumed that the two received lines have been stored in some sort of storage device. In the upper diagram, which shows the method of linear interpolation, these lines are scanned by two spots, both running with the same speed v, and the intensity I in the missing line is approximated by the arithmetical mean of the two intensities I_1 and I_2,

$$I(x) = \tfrac{1}{2}[I_1(x) + I_2(x)]$$

This is perfectly satisfactory for gently shaded areas, and also for vertical edges, but if the contour is slanting, a stepped amplitude profile arises, which, as will be shown later in practical examples, falls far short of achieving the purpose.

In the method of contour interpolation the two spots run with equal speeds v until one of them reaches an edge, i.e. a position where the intensity changes at more than a certain predetermined rate. Here it stops, while the other spot, which has not yet reached the contour, moves on with a speed $2v$, so that the interpolated spot, midway between the two, moves on with constant velocity v. This continues until the second spot has also reached the contour, after which the retarded spot gradually catches up with the other, until both move again with the same speed v. This is the kinematic rule. The intensity rule is that the interpolated intensity I is formed as the arithmetical mean of the two intensities, but *weighted with the velocities*, i.e.

$$I[\tfrac{1}{2}(x_1 + x_2)] = \frac{v_1 I_1(x_1) + v_2 I_2(x_2)}{v_1 + v_2}$$

It may be noted that the denominator is constant, as $v_1 + v_2 = 2v$. It can be seen that this rule is equivalent to linear interpolation so long as the two spots move together with the same speed, but at the instant at which the first spot is arrested the information is transferred to the one which now moves with double speed. The interpolated edge then appears at the position at which it would be if the contour in question were straight, which is the best estimate one can make in 2-point interpolation.

* We use the nomenclature in which a 'frame', which we use synonymously with e', consists of two interlaced 'fields'. This terminology has the advantage that not be misunderstood on either side of the Atlantic.

The method breaks down, of course, if the contour is so nearly horizontal that one of the scanning spots misses it. In that case an instruction must be embodied in the device to the effect that, after a certain maximum searching time, the two scanning spots must equalize their position.

The principle has been explained in its application for constructing the missing field, but it is evident that it can be equally well applied to the interpolation of missing frames. In that case the upper and the lower lines in Fig. 1 represent the same line as in the interpolated frame. There is again a good statistical reason for adopting this method of interpolation, based on the common observation that most of the changes in our world are due to moving objects, and most movements are horizontal. The upper and lower intensity profiles, $I_1(x)$ and $I_2(x)$, represent the positions of the outline of a moving object, and the assumption is that the motion can be well enough approximated by linear displacement, i.e. constant speed of motion.

In its application to frame interpolation the method will break down if the motion is so rapid that the second scanning spot will not find the edge in the allowed searching time. Assume, for instance, that we allow a searching time corresponding to 10 picture points, which is about 1/50 of a line. If we leave out one frame in two, this means that the time interval between them will be 1/12·5 sec, i.e. objects must not traverse more than one-quarter of the whole picture width in 1 sec. This is no serious limitation, because movements faster than this will appear jerky even in ordinary transmission, with 50 fields/sec. A more serious limitation is that the interpolation method does not operate in the vertical direction. It is true that by far the greater part of movements is horizontal, but the movements of the mouth form an important exception. It therefore appears likely that 12·5 fields/sec (15 fields/sec with U.S. standards), i.e. a compression ratio of 4 : 1, will be the limit to which one can go without appreciable impairment in picture quality.

There may be applications in which the advantages of reduced bandwidth will outweigh the disadvantages of some jerkiness in vertical movements. In that case one could well go to a compression ratio of 8 : 1. Fig. 2 shows two ways in which only one field in eight is transmitted, and the seven missing fields are reconstructed. In Fig. 2(*a*) this is done by repeated midway interpolation. As soon as field 1 is received, field 2 can be constructed by contour interpolation between two consecutive lines. The interpolation of the others must wait until field 9 is received. From this and field 1 first the field 5 is constructed,

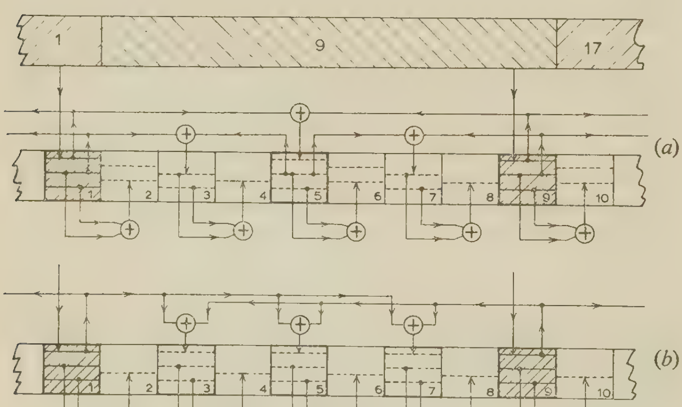


Fig. 2.—Scheme of television transmission systems with 8 : 1 compression.

(*a*) By repeated mid-way interpolation.
(*b*) By proportional interpolation.

by interpolation between their corresponding lines, and th fields 3 and 7, by interpolating between fields 1–5 and field respectively. The even fields are constructed by interpo between the lines of the odd fields.

Fig. 2(*b*) shows the somewhat simpler method of 'propor interpolation. Corresponding lines in fields 3, 5 and reconstructed from those received in fields 1 and 9 by assu that the movement of the edges was linear. This mean if, for instance, the scanning spot on line 1 stops and th line 9 moves at double speed, the writing spot on line have to move at half speed. The intensity on this line wil to be the sum of I_1 and I_9, weighted in the ratio 3 : 1. writing spot moves at half-speed the total intensity m halved, so that the previous intensity rule now changes to

$$I_3\left(\frac{3x_1 + x_9}{4}\right) = \frac{1}{8}[3I_1(x_1) + I_9(x_9)]$$

However, if the spot on line 9 stops first, the spot wil with 3/2 normal speed, and thus the factor at the righ have to be 3/8 instead of 1/8. The rule for line 7 will be similarly modified (with 1 and 9 interchanged), wh rule for line 5 is the one for midway interpolation. The struction of the even frames 2, 4, 6 is as before.

It is important to realize that contour interpolation, w between fields or between frames, taps a source of redun quite independent of the one utilized in line-by-line equalization methods, in which correlations between po one line are exploited. Combinations of the two s therefore result in a multiplication of the band-saving If, as is likely, line-by-line equalization methods are capa yielding a saving of 3 : 1 and contour interpolation savi 4 : 1 to 8 : 1, combination of the two systems will be ca of producing compression ratios of 12 : 1 to 24 : 1. Th figure comes near to one-tenth of the value which we previously estimated as the ultimate limit of all band- methods.

(4) REALIZATION OF CONTOUR INTERPOLATION PHOTO-MECHANICAL PICTURE TRANSFORMER

It was decided to demonstrate and test the principle low speed, realized by mechanical-optical means, in a 'p transformer' in which every second line of a photogra scanned, and the missing line is reconstructed by contour polation. This choice was dictated by the available func construction of a full-speed television compression system have required a considerable capital outlay for the purch storage tubes. The picture transformer has, however, a c interest in itself, as it could be used for a 2 : 1 saving in width or time for pictures transmitted by radio or by cabl

(4.1) Mechanical and Optical Design

A complete drawing of the picture transformer is sho Fig. 3. The casing forms two light-tight compartment one at the left contains the transmitter, and the one right the receiver. Running through both is a hollow s which carries a picture drum for reading at the left, and drum for recording at the right, both 60 mm long and in diameter. While the rotation is imparted to the dru the hollow spindle, the longitudinal motion is derived f rod inside the spindle, which engages the drums by pegs p through longitudinal slits in the hollow spindle. At its hand end this rod is threaded with a pitch of 0·3 mm and e with a stationary nut. Both the reading and recording are therefore scanned with 200 lines on their length of 6 and as the reading and writing spot sizes are adjusted to 0·1

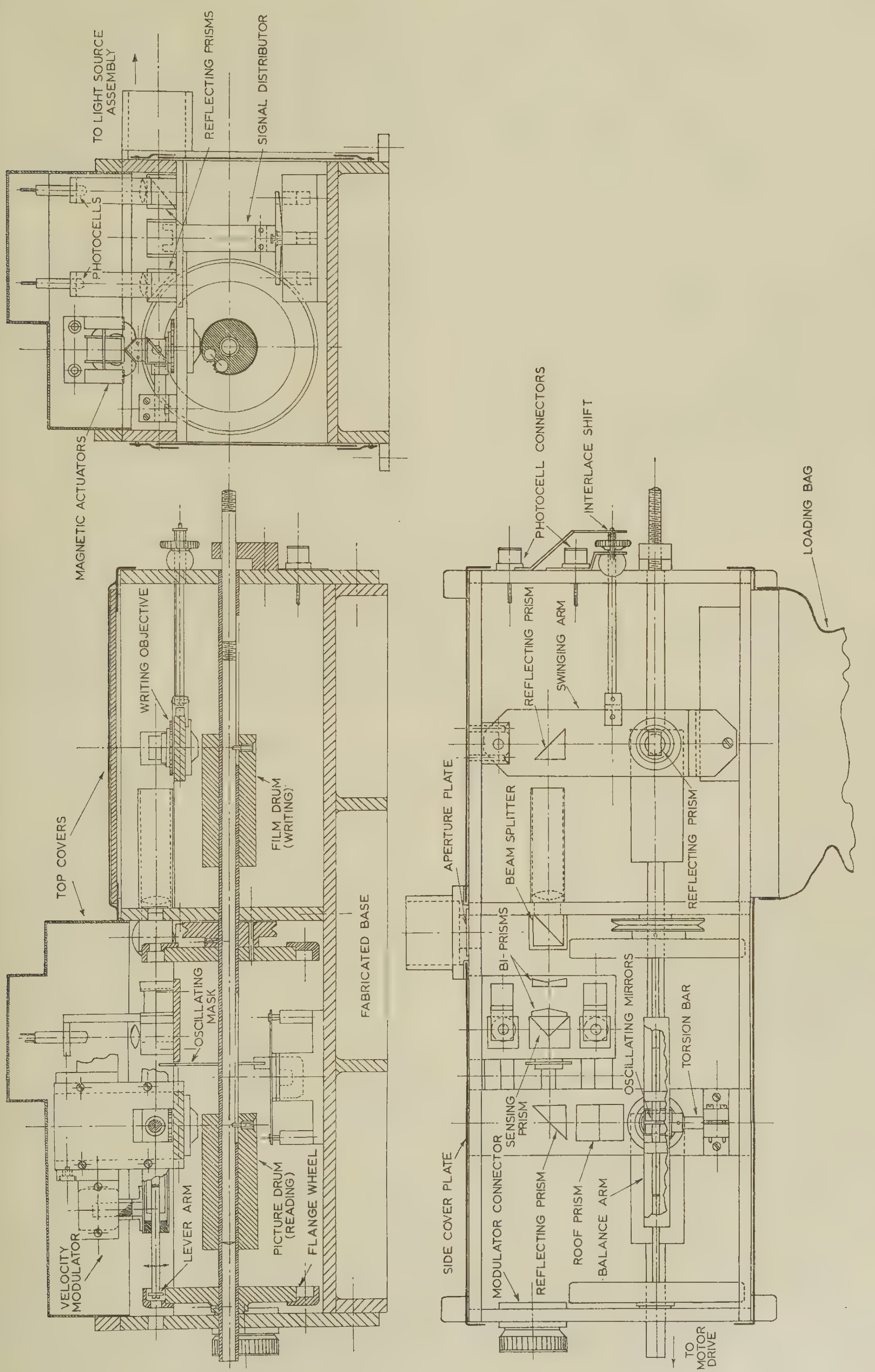


Fig. 3.—Picture transformer for 2 : 1 compression by contour interpolation.

this means scanning every second line in a photograph which measures 60 × 70 mm.

The optical system is shown in Fig. 4. The light source, which is a 150 c.p. lamp, illuminates through a condensing lens an aperture plate with two holes. The lower one serves for the main beams, for reading and writing, while the top hole is the source of the pilot beams, shown dotted, which actuate the controls. These beams start from the holes as conical bundles, but in order to simplify the representation only two rays are shown in each. They first go, as one bundle, through the lozenge-shaped signal-distributor aperture,

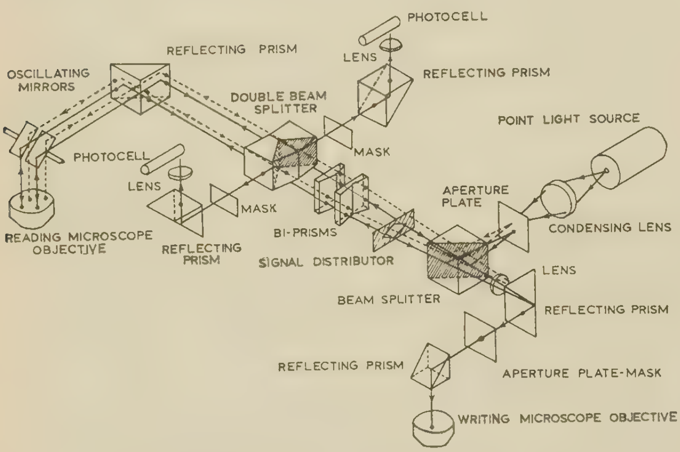


Fig. 4.—Optical system of the picture transformer.
—— Main beams. – – – Control beams.

whose function will be explained later. They are each split into two partial beams (i.e. four partial beams altogether) by a pair of biprisms, which separate the beams and collimate them. The two control beams pass through a beam-splitter prism, while the two main beams pass underneath it. This beam splitter operates usefully only on the returning beams; in the first passage it causes only an unavoidable loss of light in the control beams. After this, all four partial beams pass through a reflecting prism, which throws them on two oscillating mirrors at nearly 45° to the horizontal, and from these into a microscope objective. This is a 30 × 0·5 NA objective with 6 mm focal length, which produces on the reading drum four small images of the original apertures, all 0·15 mm diameter, arranged in a square of 0·30 mm side length. This means that the reading spots are scanning two lines, with one missing, and the pilot spots are the same distance ahead of them.

The reflected light now returns through the optical system, along the path. A fraction of the returning control beams, which report the density of the photograph at the position of the pilot spots, is split off by the double beam-splitter prism, and thrown on two photocells. The two main beams return through the aperture of the signal distributor, and are united in a spot at the right, which is the mirror image of the original aperture with respect to the beam-splitter prism at the centre. This spot, which is of the dimensions of the original aperture, contains a mixture of the two beams, in the proportion as they pass through the signal distributor aperture. It is imaged by means of a lens and two reflecting prisms through a microscope objective on the photographic film. As the spot is nearer to the film than the original aperture is to the scanned photograph, at this side we have used an objective, 40 × 0·7 NA, 4 mm focal length, so as to obtain a writing spot of the same size (0·15 mm) as the scanning spots.

As the intensity in the writing beam may well be as as 1/100 or even 1/1 000 of the original beam intensity, it the greatest importance to avoid reflections in the optical sy on the transmitting side. Blooming of the optical surfa quite insufficient. We have eliminated reflected images c illuminating aperture by rigorously avoiding optical surfa right angles to the beam. Though, for simplicity, the refl prisms are shown as 45°–90°–45°, the angles were in r 48°–84°–48°, the beam splitter is not a square but loz shaped, and the biprisms are slightly skewed. Otherwis reflected light would have drowned the picture signal i darker parts of the photograph.

The system as described, with the signal distributor placed symmetrically, would produce a 50/50 mixture of th scanned lines, i.e. it would produce linear interpolatio becomes a contour interpolating system by virtue of c accessories and controls which will now be described.

(4.2) Control System

A block schematic of the system is shown in Fig. 5, sl generalized, so that it applies also to the case of conti transmission. In the general case the input store need no more than the previous line, while in our picture transf this is the photograph, which stores all lines. Two lin scanned, each in two places. The main scan is shown in s lines, and the pilot scan in thin lines. In the picture trans the scanners correspond to the two oscillating mirrors on the reading-microscope objective. The pilot spots repo image amplitudes to contour detectors which measure th of change, and if this exceeds a certain critical level the come into action. Two actions must be performed. One change in scanning velocities, so that the spot which has a contour stops, while the other starts at double speed.

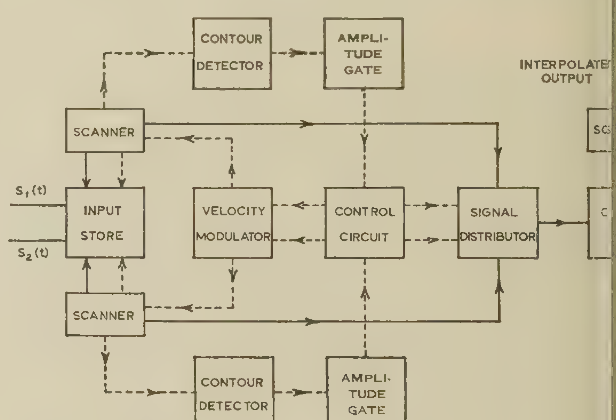


Fig. 5.—Block diagram of a 2 : 1 contour interpolation syst
—— Main channel. – – – Control channel.

effected by a velocity modulator, which in the case of the p transformer is an electromagnetic actuator, to be des later. The other action is a transfer of the signal to th which has not yet found the contour, and scans at double This is done by a signal distributor; in the picture transf this is the lozenge-shaped aperture in Fig. 4. It is moved electromagnet, which is also controlled by the control c The signal, with the properly weighted components, is put output store. In the picture transformer this is the graphic film. In the case of continuous transmission this be a transient store, scanned with an appropriate delay.

(4.3) Velocity Modulator

velocity modulator is shown in Fig. 6. Its purpose is ;st one of the spots, i.e. to let it move with the photograph, n as the corresponding pilot spot reports a contour, and ke the other move with double speed. This means that ı equal and opposite angular velocities must be applied two mirrors, which normally are stationary, at 45° to the ıntal. The time available for this acceleration corresponds delay between the pilot spot and the scanning spot, and he order of 1/100 sec.

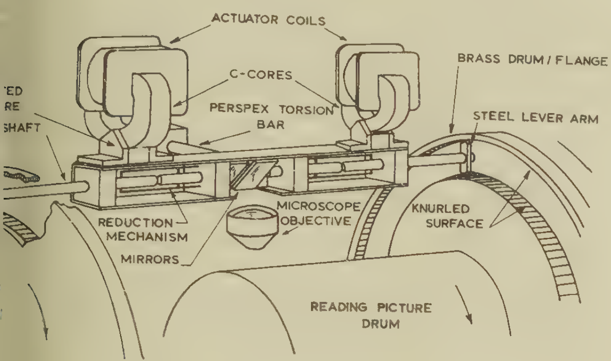


Fig. 6.—Velocity modulator.

ɛ mirrors and the steel shafts which carry them are arranged ɔalance arm, made strong and light, out of a steel channel aluminium bearings. This is held by a Perspex torsion ;o that the balance arm can be slightly tilted against the ontal. The resonant frequency was about 120 c/s and the ations were almost aperiodically damped owing to the ırable properties of Perspex.

line with the steel shafts which carry the mirrors, and which ıade hollow, are two longer steel shafts, which penetrate them so as to ensure good flexural stability. They carry eir ends steel levers, in the form of narrow segments of a ar disc knurled at the outside. These discs have a slightly er diameter than the width of the circular slots in the brass es at both ends. The brass flanges are fixed on the hollow le which carries the drum, i.e. they revolve with it but do ake part in its axial motion. The inner and outer faces e circular slots are knurled to ensure good friction. A small tilt of the balance arm corresponding to a movement out $\pm 0{\cdot}005$ in will engage one steel lever with an inside s R_1, the other with the outside radius R_2, of the brass es. Thus the levers will start turning with angular speeds $/r_1$ and $\pm R_2/r_2$ times the angular speed ω of the main where r_1 and r_2 are the radii of the levers, inside and le. We make $R_1/r_1 = R_2/r_2 = R/r$. The spots will then isplaced relative to the drum surface with velocities $R/r)\omega$, where f is the focal length of the objective. (The 2 originates from the doubling of angle at the reflection lting mirror.) Let R_D be the radius of the picture drum. If ant to arrest one of the spots, we must have

$$2fR/r = R_D$$

r case $f = 6$ mm, $R = 35$ mm, $r = 5$ mm and $R_D = 12{\cdot}7$ mm, ience the left-hand side is about $6{\cdot}5$ times larger than it to be. We had therefore to introduce an angular reduc-between the outer steel shafts and the mirrors. This anism (which can be also considered as a torque amplifier) sts of two steel spring wires, parallel to the shaft, and two wires connecting them, one engaging with the outer shaft and the other with the inner. By properly choosing the ratio of their distances from the clamping point the required angular reduction could be very exactly realized, without any sacrifice in the frequency response of the device.

The angular reduction is also convenient, because without it the excursions of the steel levers would be rather microscopic, and the criteria for the allowable slip between them and the knurled brass surfaces might be too exacting. With an objective of $f = 6$ mm and a spot size of $0{\cdot}15$ mm the mirror tilt per spot width is only 1/80, or 43′. With an angular magnification of $6{\cdot}5$ this becomes $0{\cdot}08 = 4°\,4'$, which corresponds to $0{\cdot}4$ mm at the end of a 5 mm lever. Thus an error of not more than one spot width can be realized with a knurl of a period of $0{\cdot}4$ mm, or less. If one allows a maximum searching interval of ten picture points, the maximum angular excursion of the levers will be $\pm 5 \times 4°\,4' = \pm 23°$. This searching time is sufficient for all contours having an inclination of more than 1 : 5 to the horizontal.

The balance arm, stabilized by the rather rigid Perspex torsion bar which gives it a resonance frequency of 120 c/s, requires appreciable forces for its actuation. After some trials with piezo-electric benders we turned to the magnetic actuators shown in Fig. 6. The balance arm carries near its ends two roof-shaped armatures of laminated Stalloy. These fit into the wedge-shaped gaps of C-cores of grain-oriented silicon steel, each with 1 500 magnetizing turns. The current required for the tilt is 50–60 mA.

(4.4) Signal Distributor

If there is no edge in view, the interpolated signal is a 50/50 mixture of the two intensities I_1 and I_2. This is achieved by an aperture, which can be, for instance, lozenge shaped, as shown in Fig. 4, and which cuts out one-half of each beam when it is in the symmetrical position. As soon as the contour interpolation mechanism gets into action the information must be transferred to the moving spot, while the signal derived from the arrested spot must be cut out altogether.

Our original intention was to reduce the inertia of this organ to a minimum by making the mask from a piece of photographic film, held in equilibrium by two light rubber springs, and to move it to the right or to the left by friction wheels, permanently rotating in opposite direction, and brought into contact with the film alternatingly by means of piezo-electric actuators. This is similar to the system used in magnetic-tape machines for computer stores, which can accelerate the tape to full speed in 2–3 ms, but we found that it requires a very high degree of workshop accuracy for its efficient realization. At the speeds at which the velocity modulator responded satisfactorily, a much simpler magnetic device, shown in Fig. 7, was quite sufficient.

A mask cut into a T-shaped aluminium strip is fixed at the base to a torsion organ, formed by a steel strip. Apart from being able to tilt, this strip can also move a little up or down. It carries a 3-section armature, with small Swedish iron pieces attached to a light frame (a ferrite armature might be better). Three silicon-steel C-cores, wound with about 1 500 turns, are fixed below these in an Araldite block. When the central core is magnetized the armature is made to sit down on it, thus forcing it to return to the normal position faster than if this were left to the restoring force of the steel torsion piece. When the control mechanism reports an edge, the central core is demagnetized and the armature is made to tilt towards one side or the other. The amplitude is limited in such a way that the unwanted beam (corresponding to the arrested scanning spot) is just cut off. A switching time of 5 ms was experimentally

realized; this is quite sufficient in view of the limitations of the velocity modulator. The windings of the side cores are directly in series with the windings of the magnets, which actuate the velocity distributor.

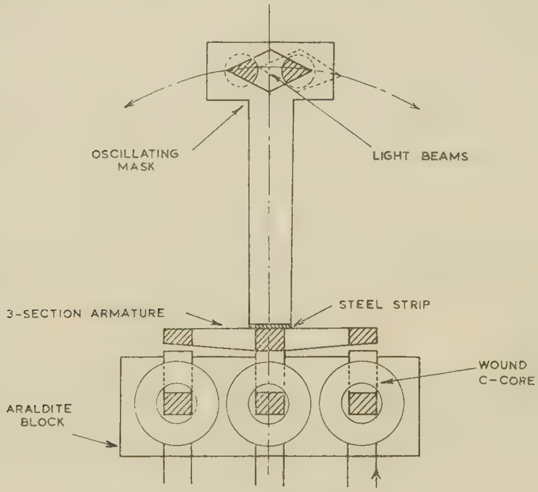


Fig. 7.—Signal distributor.

signals are further amplified, limited, differentiated and cli[p] provide a clean negative spike train indicating the leadin[g] of transitions. The circuit is provided with a cathode-f[ollower] output. The channels are provided with separate pictu[re] controls to compensate for unequal optical attenuations [in] control tracts, and the contour threshold controls are together.

The tri-stable circuit operates in a way dependent up[on] presence (or not) and on the inter-channel time displacem[ent] trigger pulses received from the contour sensing circuit circuit must be bisymmetrical in the unstable state; the ar[rival of] a pulse on either channel indicates a detected contour, a following pulse, arriving on the other channel within a period $T < T_{max} = \Delta x_{max}/2v$, indicates the other end contour successfully found by the velocity modulator s[earching] at velocity $2v$. Searching must cease if no pulse arrives [in] the preset period T_{max}. The operation may be conve[niently] condensed into a logic table.

In Table 1, Δt is the picture-element period for scanning at velocity v.

The realization of the tri-stable circuit is shown schem[atically] in Fig. 8. Two direct-coupled monostable multivibrat[ors] cross-connected to each other and also to separate gate[s] associated with the conducting valve of each pair. Th[e]

Table 1

Line 1	Line 2	Contour	Output 1	Output 2
1. No pulse	No pulse	Zero	Zero	Zero
2. Pulse, $t = 0$	Pulse, $t = 0$	Vertical	Zero	Zero
3. Pulse, $t = 0$	No pulse up to $t = T_{max}$	Horizontal slope $> -\Delta t/T_{max}$	Rectangular pulse, period T_{max}	Zero
4. Pulse, $t = 0$	Pulse, $t = T$ $T < T_{max}$	Horizontal slope $-\Delta t/T$	Rectangular pulse, period T	Zero
5. No pulse up to $t = T_{max}$	Pulse, $t = 0$	Horizontal slope $< +\Delta t/T_{max}$	Zero	Rectangular pulse, period T_{max}
6. Pulse, $t = T$ $T < T_{max}$	Pulse, $t = 0$	Horizontal slope $+\Delta t/T$	Zero	Rectangular pulse, period T

(4.5) Electronic Circuits for the Picture Transformer

The function of the circuits is to pulse drive the picture-transformer modulators according to the contour-searching routine; the inputs to the circuits are derived from the optical control tract via two photocells and are next-but-one line picture signals; the outputs provide switching instructions to the velocity modulator and signal distributor.

The circuit function may be split into distinct parts. The first detects contours of preselected 'importance' and provides two separate trigger pulse trains indicating the positions of contours as found by the pilot beams. This may be called the 'contour sensing circuit'. The second part contains the logic of the contour interpolation mechanism in a 'tri-stable circuit' which is fed by the above trigger trains, and this is followed by a modulator power drive circuit.

The contour sensing circuit comprises two identical channels and is conventional. After some experiments with gas-filled photocells and phototransistors we adopted photoconductive cadmium-sulphide cells as sensing elements.* The photocell signals are amplified and differentiated. After further amplification and filtering to remove the out-of-band high-frequency noise, the contour signal is full-wave rectified and amplitude-gated by a paraphase amplifier and diode circuit. The gated

* These had the advantage of high-level output, good thermal stability and small size, with a response time quite satisfactory for our requirements. We are indebted for them to Dr. M. E. Haine, Director, A.E.I. (Woolwich) Research Laboratory.

are simple double-triode amplitude selectors, and the[ir] waveforms are direct coupled into them from cathode l[oads of] the non-conducting halves of the multivibrators. The[se] delays are necessary to ensure that the gates remain in t[heir] state over the trigger-pulse period. In practice it was [found] that a small capacitor placed across each cathode load pr[ovides] adequate delay. The preset pulse period, T_{max}, of each [multi]vibrator is variable in the conventional way.

The operation is as follows: Assume the circuit to b[e in the] monostable state with both gates open and the multiv[ibrators]

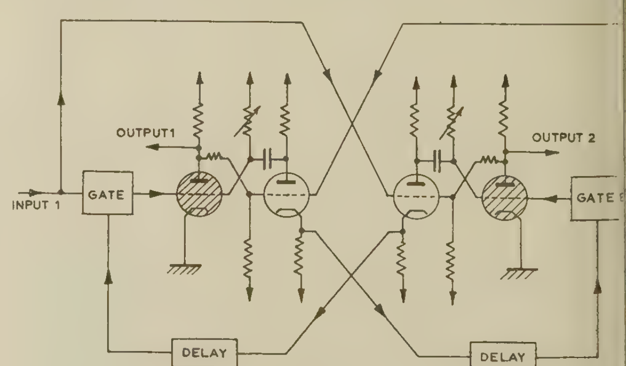


Fig. 8.—Tri-stable circuit for contour interpolation.

's conducting as shown in Fig. 8, a negative trigger pulse is 'ed on line 1 (say) and is passed by the gate, thus initiating ming action of multivibrator 1 and also closing the second The multivibrator 2 is unaffected by the arrival of the r pulse at its non-conducting valve. The action of the first vibrator then lasts until either

ı) The full quasi-stable period T_{max} of the first multivibrator is ıpleted, during which time no trigger pulse is received by line 2, the circuit then flips back into its quiescent state, or

) A trigger arrives on line 2, within a time $T < T_{max}$; this is ked by the closed gate but extinguishes the action of the first tivibrator.

ction is similar if the initial pulse is received by line 2 since rcuit topography is symmetrical about a mid-vertical line.

tput pulses from the tri-stable circuit are clipped by pentode rs and used to switch on two miniature beam tetrodes ı drive the velocity modulator and signal distributor coils ly in series. The central coil of the signal distributor is n by a third power valve switched off by either channel via miters.

(5) RESULTS OBTAINED WITH THE PICTURE TRANSFORMER

e picture transformer has been used to test and demonstrate rinciples of contour interpolation with stationary pictures; pparatus forms pictures of 'full' vertical resolution from ıalf of the information.

addition, the device was used to produce pictures containing ıll number of original scanning lines so that the quality of nterpolated results may be judged subjectively against lated full-bandwith pictures.

(5.1) Experiments

e choice of patterns and pictures for processing is an rtant matter. A good picture for the present application ins a wide range of detail in addition to well-defined large- itude contours of low geometrical slope; in addition it must pical of normal television programme material.

ree pictures have been selected:

a) A portion of a hand-drawn test chart.
b) A single black-to-white edge of medium slope.
c) An enlarged print from a portrait.

processed pictures, recorded on '120' size film, have been ged to quarter-plate size on bromide papers of carefully ed contrast grades to ensure that comparison records of the picture are not impaired by unlike ultimate photographic ties.

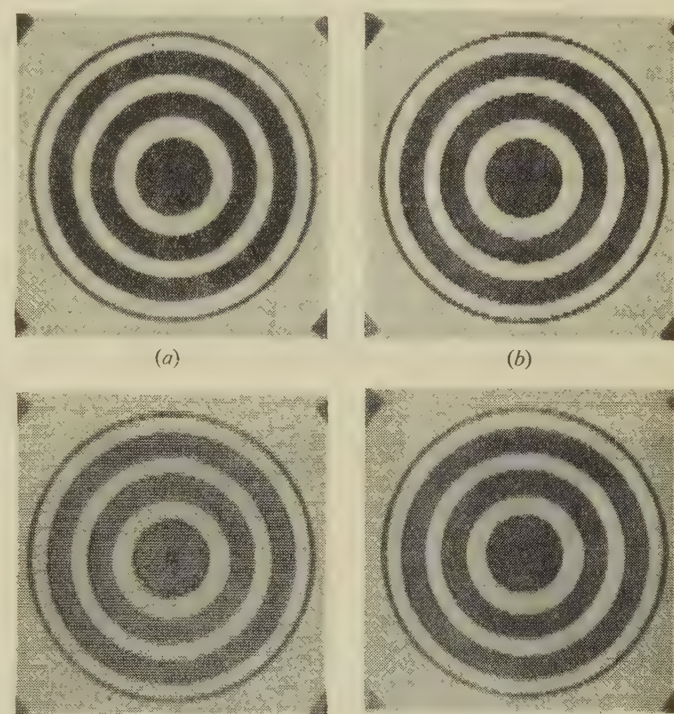
ntour-interpolated frames, in which the original 200-line is interlaced with the interpolated lines, were produced by acing, before or after the interpolation, the writing objective e line-spacing, and by moving the signal distributor mask e side, so that only one line is copied, and this with full sity. Experience has shown that it is necessary to put the line exactly between those previously recorded. Even a l overlap on one side or the other will give the impression the picture has 200 lines instead of 400. The slightest axial le or backlash in the drum drives will produce an inferior t, and much time was spent in eliminating these. If there erlap, i.e. if the spot spreads beyond the line limits, care be taken in adjusting the intensity in the second run, ıse the blank parts of the film are pre-sensitized by the first sure.

ıearly interpolated frames were taken in the same way, ɔt that the control mechanism was put out of action.

Field-repeated frames were produced by using again one reading spot, i.e. one beam only, and repeating the copying of the line with two positions of the writing objective.

The contour-interpolated records were obtained with a film reading and writing speed of 16 mm/sec, corresponding to 100 picture-points/sec or a bandwidth of about 50 c/s. Linear interpolation and line copying could be carried out at twice this speed or more.

The experimental results are shown in Figs. 9–11, and it will be observed that contour-interpolated edges are shown in one



(a) (b) (c) (d)

Fig. 9.—Test pattern with four types of processing.
(a) Full bandwidth.
(b) Field repeated.
(c) Single field.
(d) Linear interpolation.

(a) (b)

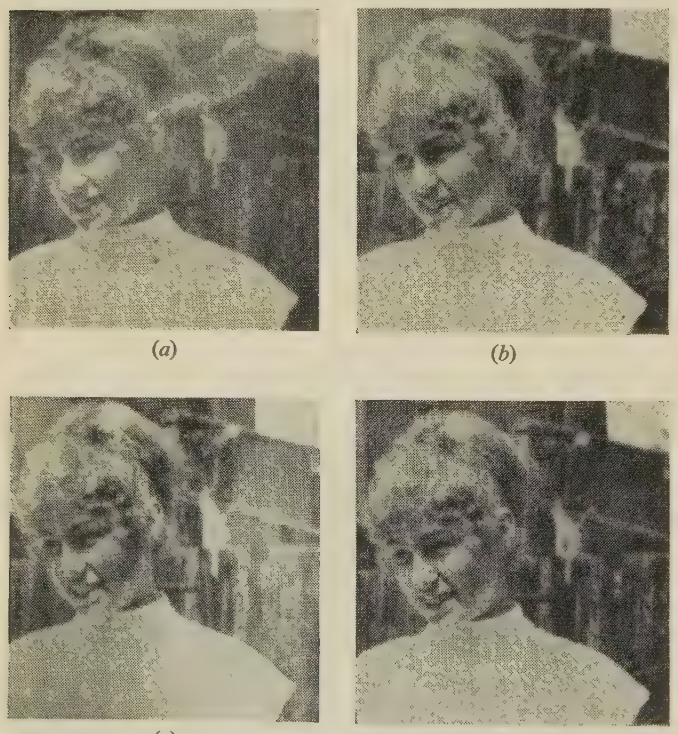
Fig. 10.—Reproduction of an edge from one field, by the missing field produced at (a) by contour interpolation, and (b) by linear interpolation.

geometrical sense only; this was due to a mechanical weakness of the velocity modulator. It was found impossible in the available time to balance the modulator arm in the quiescent position owing to the inaccurate centring of the rotating drum/flange system and also to the flexing of the balance arm and torsion bar; the result was that the modulator scanned accurately and without spurious mechanical triggering in one sense only.

(5.2) Discussion of Results and Conclusions

The enlarged interpolated edge of Fig. 10 clearly shows the transition from linear to contour interpolation, and the gain in edge sharpness by this method over linear processing is evident. This result also indicates the accuracy achieved in the instrumentation of the picture-transformer optics and modulation routine.

The assessment of the results may be left to the reader, but a few features are indicated. The single field record of Fig. 9(*c*) is interesting in that it shows how well the human eye can interpolate the missing information at the transitions; the effective contrast is reduced as the interlaced field is replaced by unexposed space. With such a test pattern, field repetition is very objectionable and linear interpolation represents only a small improvement, by smoothing out the stepped edges of sloping contours.



(*a*) (*b*) (*c*) (*d*)

Fig. 11.—A portrait processed in four ways.
(*a*) Full bandwidth.
(*b*) Field repetition.
(*c*) Contour interpolation.
(*d*) Linear interpolation.

The processes demonstrated in Fig. 11 are more revealing; field repetition produces objectionable steps in the large-amplitude transitions (the shoulders), and even the fine detail (hair) takes on a mosaic appearance and becomes slightly disturbing to the eye. Linear interpolation certainly renders the fine detail more natural compared with the field repetition, and the contours are smoothed. The contours are correctly pre in the contour-interpolated result, which shows a marke in edge sharpness. The true standard of comparison is ag 'full bandwidth' record.

The results shown demonstrate some possibilities in re processing of stationary pictures which are compressed It is regrettable that the time available did not allow the col of more material and the elimination of some instrumental nesses, but the authors believe that the results give at strong suggestion of the superiority of contour interpolatio other methods.

(6) PROPOSED ELECTRONIC REALIZATION O TELEVISION COMPRESSION

Full-speed realization of band compression by interpo will be shown to be possible utilizing conventional dev functional elements of the system, without having to d special devices such as multi-gun charge-controlled s tubes.

(6.1) Stores

At the transmitter end a single store can be used which case of $n:1$ compression simply selects one field in 'stretches' it to n conventional field periods by slow read

The storage problem is more complex at the receive must handle the stretched fields. For full-speed process re-transmission is delayed by a minimum of $(n-1)$ conve field periods. The choice of receiver storage devices depend on the following considerations:

(*a*) It is very advantageous to store and process fields fo polation in one store; this avoids accurate store match necessitates stores of twin field or full frame capacity.

(*b*) Field interpolation requires each line of the received be available three times, once for its own full-speed repetit twice for interpolation of lines one above and one below i interlaced field.

(*c*) In frame interpolation each received line is required o interpolation of the corresponding line in the previous fram for its own reproduction, and again for interpolating the fo frame, in that order.

The ideal for this application is therefore a half-tone of frame capacity with two writing organs, two indep reading organs and controllable erasing action for m read-outs, with an erase period less than the conventional blanking period.

We believe that at the present state of the storage-tu the most convenient arrangement is provided by a came optically coupled to a double-beam cathode-ray tube problems encountered in facing camera tubes with catho tubes are well known in standards conversion[24] and will elaborated.

(6.2) Contour Interpolation with a Single-Beam Cam Tube

We assume that a full frame of lines has been written camera mosaic. In the case of field interpolation this is recorded as an odd field interlaced with itself, both simultaneously. For frame interpolation the frame sequential interlace of an odd field with the next-but-one

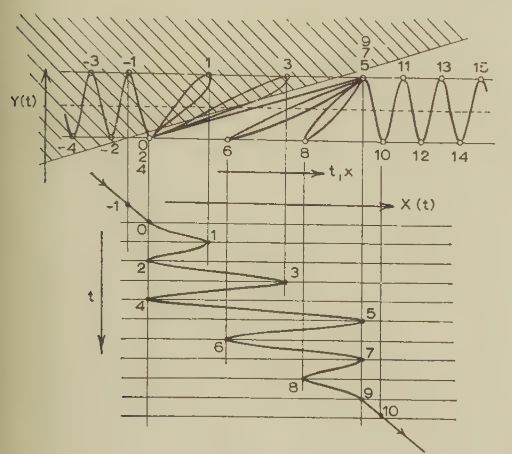
It will now be shown how such a frame may be read c lines at a time by vertically oscillating a single reading b sample them at the Nyquist rate. A method for modulating this oscillating scan is illustrated in Fig. the spot wobble superposed on the Y-scan is given by

$$y(t) = d \sin 2\pi f_s t$$

:e f_s is the sampling frequency and d is half the line spacing, nodulation is effected by augmenting the X-scan by a sinus- oscillation $x(t)$ in phase with $y(t)$ and amplitude modulated a ramp function of slope equal to the normal horizontal ıing velocity v, with a polarity dependent upon the slant of ontour, i.e.

$$x(t) = \pm vt \sin 2\pi f_s t$$

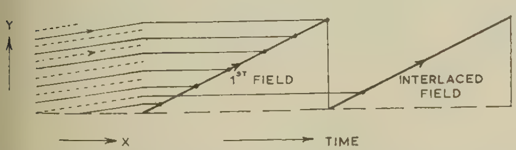
the onset of the contour occurring at $t = 0$, so that $= 0$, for $t < 0$. These oscillatory scans combine to give $= \pm vty(t)/d$, and so along the uppermost line we have



12.—Scanning scheme for interpolation between two lines with one reading spot.

$= \pm vt$ and along the lower line $x_2(t) = \mp vt$, superposed e main scan of $X(t) = vt$, which gives the required velocity ılation. The ramp function is made to run down to zero e same rate once the contour is successfully found.

an alternative, the velocity modulation may be performed de the camera tube by delay-line techniques. The mecha- would be as follows: each sampled line signal derived from ra spot wobble is propagated down a separate parallel bank lay lines of delay nT, where T is the picture element period ıs approximately) and n ranges down the banks from 0 to V being the maximum number of 'search elements'. Linear polation is given by the arithmetic mean of the outputs ed from the central delay lines (delay NT) of the two banks, he velocity modulation is performed by switching across vo sets of delay-line terminations, say at 3 Mc/s, in opposite ions.



3.—Simplified representation of scanning, in which each field is represented by a single slanting line.

(6.3) 2 : 1 Compression Systems

field interpolation the transmitter requires a single store ich only odd fields are stored and scanned at half speed. s achieved by blanking out alternate fields on a cathode- ıbe displaying the conventional signal, faced with a low- camera tube. The transmission channel handles stretched fields of half the conventional bandwidth which constitute the receiver input.

A receiver which reconstructs the sequential interlaced video signals is shown schematically in Fig. 14. In this Figure, A is a tube of field capacity with a single writing gun and one reading gun with 'read/erase' action (e.g., an image orthicon facing a cathode-ray tube). B is a store of frame capacity with a

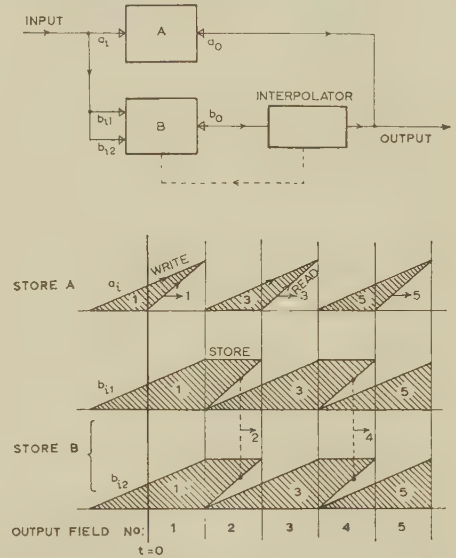


Fig. 14.—Scheme and scanning diagram of a 2 : 1 field interpolator.

double writing gun and single reading gun with read/erase action (image orthicon facing a double-beam cathode-ray tube). The scanning sequence of the two stores is also shown; the diagrammatic representation of scanning in which a field is represented by a single slanting line is explained in Fig. 13. The shaded heights in these diagrams indicate the proportions of fields stored at any time t.

The receiving process is as follows: commencing with A and B empty, let the field 1 be half received and stored on A and also, in duplicate, on B at time $t = 0$. This field is now read out from A directly at full speed at the same time as the second half continues to be stored on A and on B at half speed. The interlaced field 2 is now formed by reading out B with contour interpolation. During this period field 3 is half stored by A and B at half speed, thus completing the cycle of two fields. The overall delay in this transmission system is one field period.

Pure 2 : 1 frame interpolation is applicable to a non-interlaced system. The transmitter sends only alternate frames stretched to occupy two frame periods. It can be shown that a receiver for reconstructing the original signal with contour interpolation requires two stores of double frame capacity, each with twin writing guns and single read/erase guns. Such a system might be of interest for industrial or commercial television, where it is desirable to double the definition in the full waveband, rather than to transmit the standard definition in half the waveband. The superiority of sequential over interlaced scanning at the same number of lines is almost 2 : 1; only flicker prevents us from exploiting it in conventional television.

(6.4) Multiple Interpolation Systems

Combination of the 2 : 1 field and frame-interpolation systems gives a 4 : 1 compression scheme in which only alternate odd fields (1, 5, 9 . . .) are selected and stretched at the transmitter.

The receiver utilizes four stores, three of which are single-gun writing and reading stores used for frame interpolating, the fourth being used for all field interpolations with a twin writing gun. The receiver thus comprises four camera tubes and cathode-ray tubes, one of which is a double-beam tube. This arrangement provides a receiver output in the following sequence:

Field 1. Full-speed replica of the received quarter-speed field.
Field 2. Interlaced field interpolated from previous field.
Field 3. Frame interpolated from received fields 1 and 5.
Field 4. Field interpolated from previously formed field 3.
Field 5. Full-speed replica of received field 5.

No output store is required.

The most ambitious scheme which one might contemplate is an 8 : 1 compression system with complete contour interpolation. The procedure for obtaining seven missing fields out of eight has already been indicated in Fig. 2. In the most economical receiver all frame interpolations (fields 3, 5, 7; 11, 13, 15; . . .) are performed at one reading of the received field pairs (1, 9; 9, 17; . . .) and intermediate storage is used for those fields which are not immediately transmitted.

A receiver for 8 : 1 compression with complete contour interpolation requires six stores (six camera tubes and six cathode-ray tubes with one of these double-gun and two with variable-velocity scanners). The scanning sequence is given in Fig. 15; A is a store for field copying, B and C are frame-interpolating stores, D serves for all field interpolations, and E

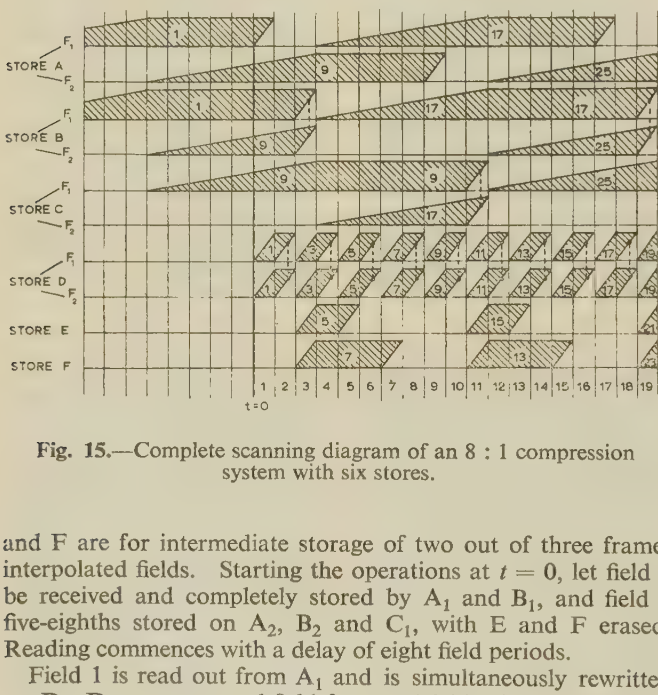


Fig. 15.—Complete scanning diagram of an 8 : 1 compression system with six stores.

and F are for intermediate storage of two out of three frame-interpolated fields. Starting the operations at $t = 0$, let field 1 be received and completely stored by A_1 and B_1, and field 9 five-eighths stored on A_2, B_2 and C_1, with E and F erased. Reading commences with a delay of eight field periods.

Field 1 is read out from A_1 and is simultaneously rewritten on D_1, D_2 as a repeated-field frame. Field 9 is three-quarters stored on A_2, B_2 and C_1. Field 1 is stored by B_1.

Field 2 is field interpolated from field 1 on D, and field 9 is seven-eighths stored on A_2, B_2 and C_1. Field 1 is stored by B_1.

Field 3 is frame interpolated from fields 1 and 9 on store B, and fields 5 and 7 which are simultaneously interpolated are written into E and F, respectively. Field 3 is fed back to D as a repeated-field frame. C_1 and A_2 fully store field 9, and A_1, B_1, B_2 and C_2 are erased.

Field 4 is field interpolated from field 3 on D. Field 17 is one-eighth received by A_1, B_1 and C_2, and A_2 and C_1 completely store field 9. E and F continue to store fields 5 and 7.

Field 5 is read directly from E and rewritten back into D repeated-field frame; field 17 is one-quarter stored by A and C_2, and F stores field 7 with field 9 remaining in A_2 an

Field 6 is field interpolated from field 5 on D_1 and D_2; fi is three-eighths stored on A_1, B_1 and C_2, and field 9 is sto A_2 and C_1; F continues to store field 7.

Field 7 is read out from F directly and fed back to D repeated-field frame; A_1, B_1 and C_2 have received half of fie and A_2 and C_1 store field 9.

Field 8 is field interpolated from field 7 on D; field five-eighths stored in A_1, B_1 and C_1, and A_2 and C_1 store fi E, F, and B_2 are empty.

The remaining half-cycle is now evident.

(6.5) Combined Compression Schemes

The main attraction of bandwidth saving by interpo lies in the possibility of combining the system with other pendent compression systems to produce a large savi bandwidth. If, for instance, a rate-equalizing system placed in tandem with the field-stretching transmitter, the would operate on conventional fields at a reduced band The receiver would then consist of a decoder followed contour interpolator. It may well be possible to make the system simpler than the sum of its two component sy because all line-by-line rate-equalizing systems contain means for locating edges. Hence it is not necessary to this operation in the contour-interpolation system. Su transmission system could have a compression ratio of 2 and with this the possibility emerges for the first time of in taneous transatlantic television via cable or short-wave link.

Such an international exchange requires standards conve and this is easily combined with the interpolation process. standards step-down transmissions, such as American (525 60 fields/sec) to British (405 lines, 50 fields/sec) the trans store selects one field in 12 and stretches it to ten converte periods with 202½ lines; the receiver then operates on a compressed signal, frame interpolating four fields and interpolating five fields between every two received stre fields. Alternatively 5 : 1 compression could be achiev transmitting one American frame in six with 405 lin 5 frames/sec. The reverse transmission involves st standards conversion; if the British 405-line 50-fiel signal is 10 : 1 compressed the receiver must interpolate and increase the field line number by a factor of 5/4. Th operation can be obviated by duplicating the initial p tube with a camera working to the higher standard and compressing 12 : 1 at the transmitter. If these compr figures prove to be unattainable, the two functions m separated and the standards convertor would work in ta with the 8 : 1 compression system in such an order th transmission channel handles the compressed lower sta signal.

(7) CONCLUSIONS AND OUTLOOK

Line-by-line rate-equalization methods promise, by them band saving in a ratio of 3 : 1, or perhaps a little better. Co interpolation by itself offers a 4 : 1 saving without appre impairment of quality, and 8 : 1 if one accepts some dete tion in the reproduction of vertical movements. In combi the two systems offer potential compression ratios of 12 24 : 1.

Neither of these systems is as yet a practical propo Both involve complicated processing, which must be carri with high precision, excluding systematic and statistical

processing apparatus. In practice, even the first modest level quantization, has led to rather objectionable picture s. Nevertheless, from a purely scientific point of view the ems are soluble, given sufficient capital and effort. Legiti- doubts arise only if one considers whether the result would omically justify the required investment and work.

appears most unlikely that even the simplest band-saving ns, such as storage tubes, repeaters with magnetic or tube ge, variable-velocity schemes or field interpolation, will ever nbodied in home receivers. The reason is not only the cost, but also the incompatibility of band-saving systems, n would make it necessary to provide new wavebands for . This difficulty does not exist in closed-circuit systems, or the time being most of these operate under conditions e there is no restriction on the waveband. There is a in demand for commercial television links over greater nces, through cables, but television *via* telephone lines is not ctical possibility, and in most cases a cable of a few hundred ycles per second bandwidth, plus the complicated terminal oment, appears to be too large an economic handicap for a service. At present we see greater possibilities in improving definition of closed-circuit television, where there is no iction of waveband by using sequential scanning with ge display tubes or with repetition or with contour inter- tion, than in systems transmitting the rather low picture ity at present acceptable in a reduced waveband.

e greatest impetus for waveband reduction might be cted to come from a desire to realize transatlantic television, e the waveband would have been greatly restricted if we ted to accommodate it in the transatlantic cable or in short- radio links. Unfortunately this impetus is not likely to be coming, for the simple reason that there is a five-hour time rence between London and New York. A jet plane which take a magnetic tape record in five hours from London ort to Idlewild would make events in London practically ltaneous for the viewers in New York, while a simultaneous smission would arrive five hours too early. Nothing could e events in the United States practically simultaneous for ers in Western Europe; the direct transmission would arrive hours too late, and there are only very few events per year for h the viewers could not as well wait until the next evening. is therefore unlikely that a project of simultaneous trans- ntic television will evoke the rather considerable funds ired for its realization. For the time being at least, tele- n compression schemes will probably be restricted to much ambitious applications.

(8) ACKNOWLEDGMENTS

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THE EFFECTS OF PARASITIC MODULATION ON THE ACCURACY OF MEASUREMENT OF THE Q-FACTOR OF A RESONATOR

By F. H. JAMES

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SUMMARY

When the Q-factor of a resonator is determined by using sinusoidal amplitude modulation and measuring the envelope phase shift, the presence of unwanted frequency or phase modulation may reduce the accuracy of the result. The envelope phase shift now depends on the ratio of the parasitic-modulation index to the amplitude-modulation index, and also on the phase angle between the two modulation components. If measurements are made with the carrier set as close as possible to the resonant frequency of the resonator, increased frequency stability is required. Alternatively, if advantage is taken of an approximation to exact tuning, which is given by setting the carrier so that the envelope phase shift is a maximum, a systematic error is introduced. If the ratio of the modulation indices or the phase relation between the modulation components is known, the systematic error may be determined and a correction can then be applied to the result. If either of two special phase relations can be set up between the modulation components, it appears that the parasitic modulation can be rendered innocuous; under these conditions as high an order of accuracy can be achieved as if pure amplitude modulation were used. This paper is supplementary to a previous paper which described a method for the measurement of very high Q-factors of electromagnetic resonators.

(1) INTRODUCTION

When a sinusoidal amplitude-modulated signal is applied to a tuned circuit, the envelope is subjected to a phase shift which increases with the modulation frequency. A previous paper[1] outlined two alternative procedures by which this principle could be applied to the measurement of the Q-factor of a resonator. Other authors have described how these procedures may be used for the measurement[2,3] of Q-factors at 700 Mc/s and 3 000 Mc/s. Whilst it may be possible to produce pure amplitude modulation, in general, frequency or phase modulation are present as parasitic modulation components. In order to determine the errors introduced into the result for the Q-factor when parasitic modulation is present, the simple analysis which was previously given must be modified. In addition to the amplitude-modulation sidebands, those due to parasitic modulation must be considered. If the parasitic-modulation index is restricted to values less than $0\cdot4$, only the first pair have any significance. One more modification is necessary; account must be taken of the phase relation between the two modulation components. The phase angle between the upper sideband component of the parasitic and amplitude modulation can have any value between 0 and 2π. Thus this whole range must be investigated for a general solution to the problem.

(2) MODIFIED EXPRESSION FOR THE ENVELOPE PHASE SHIFT

It was shown in a previous paper that, when pure amplitude modulation was used, the envelope phase shift produced by a tuned circuit was given by the following expression:

$$\gamma = \text{arc tan}\,[a_2A_0/(A_0 + a_1^2A_2)]$$

Where $a_2 = 2Q\omega_m/\omega_0$
Q = Q-factor of the circuit.
ω_m = Angular frequency of the modulation.
ω_0 = Angular frequency of the tuned circuit resonance.
$A_0 = 1 + a_1^2 + a_2^2$.
$a_1 = 2Q\omega_d/\omega_0$.
$\omega_d = \omega - \omega_0$.
ω = Angular frequency of the carrier.
$A_2 = 1 + a_1^2 - a_2^2$.

However, when parasitic modulation is present, it may be sh that the envelope phase shift is now given by

$$\gamma = \text{arc tan}\,\frac{a_2A_0 - 2a_2cA_1 + sA_1A_2}{A_0 + a_1^2A_2 + cA_1A_2 - 2a_2sA_1}\ .$$

where $A_1 = a_1a_2m_p/ma$.
m_p = Parasitic-modulation index.
m_a = Amplitude-modulation index.
$c = \cos\mu$.
$s = \sin\mu$.
μ = Angle between the upper sideband component o parasitic- and amplitude-modulation vectors.

(3) CALCULATION OF THE ACCURACY ATTAINABL

When the Q-factor of a resonator is determined by this met the carrier frequency is first adjusted to be as near as pos to the resonant frequency of the resonator, and then the enve phase shift is measured. From eqn. (1) we must there determine the variation of the envelope phase shift as a func of the carrier frequency for fixed values of the other parame The parameter a_1 represents the difference frequency betw carrier and resonant frequency, multiplied by the ratio betw twice the Q-factor and the resonant frequency. Thirteen va of a_1 were taken in order to include carrier frequencies a and below the resonant frequency up to limits represente $a_1 = \pm\,0\cdot6$. Seven different values of the modulation-i ratio were taken, from 0 to $0\cdot3$. Sixteen values of the angle μ were used to cover the possible values from 0 to The parameter a_2 represents the modulation frequency multi by the ratio of twice the Q-factor to the resonant freque Since $a_2 = 1$ when the envelope phase shift is equal to $\pi/4$ and when the carrier frequency is equal to the resonant freque three values of a_2 were taken. These values were above, b and including unity. A suitable programme was written fo C.E.R.N. Mercury computer, and the results were spe obtained as a set of tabulated values for the envelope p shift. These Tables have been used in obtaining the re given in the paper.

(3.1) Parasitic Modulation Absent

(*a*) a_2 *not Greater than Unity.*

The envelope phase shift is constant at its maximum provided that the carrier frequency does not differ from

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
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esonant frequency by more than $\pm 0{\cdot}1 f_0/Q$. If the difference etween the carrier and resonant frequency becomes as large s $\pm 0{\cdot}15 f_0/Q$, the envelope phase shift will be reduced by ·002 rad at these two limits. The relative error in the measured alue of the Q-factor will then be $-0{\cdot}4\%$. Thus if a Q-factor f 10^5 is measured at a resonant frequency of 200 Mc/s, the rror of a single measurement should not be worse than $-0{\cdot}4\%$ rovided that the carrier frequency is set to within ± 300 c/s of ie resonant frequency.

)) a_2 Greater than Unity.

There are now two equal maxima for the envelope phase shift; iey occur for carrier frequencies situated symmetrically about ie resonant frequency. For $a_2 = 1{\cdot}1$, the envelope phase shift ianges by $0{\cdot}002$ rad from its value at the resonant frequency 'hen the carrier is changed by $\pm 0{\cdot}15 f_0/Q$ from the resonant 'equency. It is possible to use these data as the basis of a alibration system for a_2 when parasitic modulation is absent. ı addition, the appearance of these two equal maxima equiɔaced about the resonant frequency indicates that parasitic iodulation is not influencing the result.

(3.2) Parasitic Modulation Present

ı) $\pi/4 < \mu < 5\pi/4$.

For this range of phase angles, the maximum of the envelope hase shift always occurs when the carrier frequency is higher ıan the resonant frequency. Let us describe this particular ırrier frequency as the 'apparent resonant frequency'. For any xed phase angle in this interval, both the apparent resonant 'equency and the envelope phase shift at this frequency increase 'ith the modulation-index ratio. For any fixed ratio of the ıodulation indices, as the phase angle increases from $\pi/4$ to $\pi/3$, once again both the apparent resonant frequency and the ıvelope phase shift at this frequency increase. However, both iese quantities decrease as the phase angle increases from $\pi/3$ to $5\pi/4$. If measurements are made of the Q-factor with ie carrier set approximately to the resonant frequency of the esonator, the errors that are introduced will depend on how ıpidly the envelope phase shift varies with the carrier frequency. 'learly this is a function of the modulation-index ratio and of ie phase relation between the modulation components. Supose now either that one is unaware of the presence of parasitic ıodulation, or that one wishes to take advantage of the maxiıum of the envelope phase shift as a more ideal point for ıeasurement, for at this carrier frequency the envelope phase iift is stationary. Since the phase shift is now dependent on ie ratio of the indices and on the phase angle between the two odulation components, a systematic error depending on these vo quantities is introduced. To give limits for the ratio of e modulation indices when a certain accuracy is desired for easuring the Q-factor, Tables 1 and 2 have been compiled.

Table 1

MEASUREMENTS MADE NEAR THE RESONANT FREQUENCY

Ratio of indices	Average error in Q-factor for a single measurement at the stated phase angle		
m_x/m_a	$\mu = \pi/3$	$\mu = 2\pi/3$	$\mu = \pi$
	%	%	%
0·05	±0·2	±0·7	±0·5
0·1	±0·3	±1·3	±1·0
0·15	±0·5	±2·1	±1·5
0·2	±0·7	±2·7	±2·0
0·25	±0·9	±3·3	±2·5
0·3	±1·1	±4·1	±3·0

Table 2A

MEASUREMENTS MADE AT THE APPARENT RESONANT FREQUENCY WHEN $\mu = 2\pi/3$

Ratio of indices	Apparent resonant frequency in terms of a_1	Average error of Q-factor apparent	Systematic error
		%	%
0·05	+0·3	−0·2	+1·6
0·1	+0·4	−0·4	+4·0
0·15	+0·45	−1·0	+6·6
0·2	+0·5	−0·8	+9·8
0·25	+0·5	−0·8	+13
0·3	+0·6	−0·8	+16

Table 2B

MEASUREMENTS MADE AT THE APPARENT RESONANT FREQUENCY WHEN $\mu = \pi/3$

Ratio of indices	Apparent resonant frequency in terms of a_1	Average error of Q-factor apparent	Systematic error
		%	%
0·05	+0·25	−0·2	+0·4
0·1	+0·3	−0·2	+1·0
0·15	+0·35	−0·5	+1·6
0·2	+0·4	−0·8	+2·6
0·25	+0·45	−0·8	+3·4
0·3	+0·5	−0·5	+4·8

Table 2C

MEASUREMENTS MADE AT THE APPARENT RESONANT FREQUENCY WHEN $\mu = \pi$

Ratio of indices	Apparent resonant frequency in terms of a_1	Average error of Q-factor apparent	Systematic error
		%	%
0·05	+0·25	−0·4	+1·0
0·1	+0·3	−0·4	+2·4
0·15	+0·35	−0·9	+3·8
0·2	+0·4	−0·4	+5·4
0·25	+0·4	−0·4	+7·8
0·3	+0·4	−0·6	+8·6

Table 1 is concerned with measurements made when the carrier is adjusted to be as close as possible to the resonant frequency. Tables 2A–2C are concerned with measurements made when the carrier is adjusted to the apparent resonant frequency. For all the Tables, the accuracy of setting the carrier frequency to the resonant frequency or to the apparent resonant frequency is taken as $\pm 0{\cdot}1 f_0/2Q$. Since there are slight differences in the error for the Q-factor when the error in the carrier frequency is positive or negative, the average of these two is quoted in the Tables.

(b) $5\pi/4 < \mu < \pi/4$.

For this range of phase angles, the maximum of the envelope phase shift always occurs when the carrier frequency is lower than the resonant frequency. For any fixed phase angle, as the ratio of the modulation indices increase so the apparent resonant frequency decreases; at the same time the envelope phase shift. increases. For any fixed ratio of the modulation indices, the apparent resonant frequency decreases as the phase angle changes from $5\pi/4$ to $5\pi/3$, and at the same time the envelope phase shift at this carrier frequency increases. As the phase angle changes

from $5\pi/3$ to $\pi/4$, the apparent resonant frequency increases, and the envelope phase shift decreases. Tables 1 and 2 may be used for this range of phase angles provided that $4\pi/3$ is substituted for $\pi/3$, $5\pi/3$ is substituted for $2\pi/3$, and 2π is substituted for π. In Tables 2A, 2B, and 2C the sign of a_1 should now be made negative for this range of phase angles. As previously stated, the envelope phase shift is always greater than that at the resonant frequency when the carrier is set to the apparent resonant frequency. The sign of the systematic error is therefore unchanged.

(*c*) $\mu = (4n + 1)\pi/4$, where $n = 0, 1, \ldots$

When any of these phase relations can be set up between the modulation components, provided that the modulation-index ratio does not exceed 0·15, the envelope phase shift varies with carrier frequency in a manner similar to the variation when parasitic modulation is absent. The differences between the two variations are always less than 0·001 rad. For $\mu = \pi/4$ and a modulation-index ratio of 0·3, the envelope phase shift is different from its value at the resonant frequency by +0·001 and −0·004 for differences between carrier and resonant frequency of $+0 \cdot 15 f_0/Q$ and $-0 \cdot 15 f_0/Q$, respectively. When $\mu = 5\pi/4$ and again the ratio of the modulation indices is 0·3, the signs and magnitudes of the errors are interchanged; thus the envelope phase shift is now different from its value at the resonant frequency by −0·004 and +0·001 for differences between carrier and resonant frequency of $+0 \cdot 15 f_0/Q$ and $-0 \cdot 15 f_0/Q$, respectively. The asymmetry in the errors is due to the appearance of the apparent resonant frequency. The change in sign and magnitude of the error as the phase angle is altered from $\pi/4$ to $5\pi/4$ is due to the apparent resonant frequency being higher than the resonant frequency for $\mu = \pi/4$, and lower than the resonant frequency for $\mu = 5\pi/4$.

(4) EXPERIMENTAL EVIDENCE

A series of measurements were made of the Q-factor of a resonator in order to test the results of the analysis and the subsequent computation. The experimental details are as follows: A stable-frequency cavity-controlled oscillator followed by an amplifier was used as the variable-frequency r.f. source. The output from an *RC* oscillator of variable frequency was used to modulate the amplifier. The frequency of the modulation and the carrier were measured by means of a commercial frequency counter. An output from the amplifier was coupled to the resonator under test. Rectified signals originating before and after the resonator were applied to a phase-difference detector, which was adjusted so that $\pi/4$ rad was indicated by a null reading on a sensitive meter. The amplified rectified signals were monitored by means of an oscillograph. No attempt was made to measure either the phase angle between the modulation components or the ratio of the modulation indices. The observations were confined to carrier and modulation frequencies for envelope phase shifts of $\pi/4$ rad. The following procedure was employed: first, the carrier was adjusted to be as close as possible to the resonant frequency, and then the modulation frequency was set to give an envelope phase shift of $\pi/4$ rad. Both frequencies were measured and recorded. Then the carrier was adjusted so that the maximum envelope phase shift was obtained; a readjustment of the modulation frequency was then made, if necessary, in order to give an envelope phase shift of $\pi/4$ rad. Again carrier and modulation frequency were measured and recorded. The Q-factors at the two carrier frequencies could then be calculated.

An adjustment was then made to the r.f. source, which had the effect of altering the phase angle between the modulation components. The previously described procedure was again employed. The results for six such sets of results are given in Table 3.

Table 3

RESULTS OF SIX TESTS

Carrier frequency	Q	a_1	Difference between two results for Q-factor
kc/s			%
*202 493·10	17 500		
202 494·81	17 720	+0·3	1·2
*202 492·96	17 550		
202 494·08	17 600	+0·2	0·3
*202 492·34	17 500		
202 490·05	17 750	−0·4	1·4
*202 492·60	17 400		
202 488·93	17 950	−0·6	3·1
*202 492·75	17 350		
202 488·95	18 200	−0·65	4·9
*202 492·84	17 500		
202 488·22	18 500	−0·8	5·7

* The measurement was made with the carrier set as close as possible to the resonant frequency of the resonator.

(5) CONCLUSIONS

The experimental evidence is all in support of the theory, although it cannot be conclusively tested since neither the modulation index nor the phase angle were measured. If either of these two quantities were measured, a fifth column could have been given in Table 3 to show how well theory and experiment agreed. If the results of the measurements which were made with the carrier set as close as possible to the resonant frequency are examined, it will be seen that the extreme values for the measured Q-factor are within 1%. If we apply Table 1 to these results, it is only necessary to set the carrier to within ±500 c/s of the resonant frequency; clearly from inspection of the measured values of the carrier frequencies, much more precision was achieved, and hence the possible fluctuations of the measured Q-factor will be much less than those shown in Table 1. The measurements made with the carrier set to the apparent resonant frequency indicate that, for at least one set of results, μ was either $\pi/4$ or $5\pi/4$ or rather close to one of these values. The result for the measurement of the Q-factor under this condition agrees very well with the results obtained when the Q-factor was measured with the carrier set to the resonant frequency. As a general conclusion, it would seem that this method could be used as a precision method for the determination of the Q-factor of a resonator.

(6) ACKNOWLEDGMENTS

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THE DESIGN OF AN AUDIO-FREQUENCY AMPLIFIER FOR HIGH-PRECISION VOLTAGE MEASUREMENT

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SUMMARY

The specification and design of an amplifier required for precision dio-frequency measurements is discussed. Formulae are presented r the 'ring of three' feedback circuit and a detailed analysis is made appropriate feedback theorems. A multi-gain amplifier which tends the voltage range of the electrostatic voltmeters, used at the .P.L. as basic a.c./d.c. transfer instruments, is described. An crease in this range of up to 1 000 times is provided and enables ltages between 60 mV and 60 volts in the frequency range 30 c/s– kc/s to be measured to a few parts in 10^4.

LIST OF PRINCIPAL SYMBOLS

A = Complex internal gain of amplifier.
A_m = Internal gain at mid-frequency (real).
$_1, A_2, A_3$ = Complex gains of stages 1, 2 and 3, respectively.
A' = Complex gain of amplifier with feedback.
A'_m = Gain of amplifier with feedback at mid-frequency (real).
A'_h, A'_l = Complex gain of amplifier with feedback at high and low frequencies, respectively.
β = General feedback factor.
β_1, β_3 = Feedback factors defining transfer from 1st and 3rd stages to input, respectively.
$\tau = CR$ = Dominant time-constant in amplifier.
n_2, n_3 = Factors by which τ is multiplied to give second and third time-constants, respectively.
Z_s = Source impedance.
R_s = Source resistance.
C_s = Series source capacitance.
Z_g = Grid–cathode impedance of input valve.
C_g = Grid–cathode parallel-capacitance of input valve.
ω = Angular frequency.

(1) INTRODUCTION

The standard a.c./d.c. transfer instrument used at the N.P.L. the electrostatic voltmeter designed by Rayner.[1] The transfer ror of this instrument at power frequencies is believed[2] to be out 1 part in 10^4, and a determination by a completely different ethod on a modified instrument at the Australian Standards boratory[3] is in agreement with this value. Subsequent published work at the N.P.L., using thermocouples and ermistors as comparison transfer elements, indicates that this ror is independent of frequency up to at least 30 kc/s and does t change by more than about 1 part in 10^4 up to 100 kc/s. is latter change is of the order to be expected from the voltage op due to the capacitance current flowing in the suspension d the measurement of the resonant frequency of the instrument. ost of the instruments at the N.P.L. were designed to be used the voltage range 65–115 volts, and have a reading accuracy approximately 1 part in 10^4 at these extremes and of approximately 3 parts in 10^5 at the optimum setting of 100 volts. Voltages above the direct range of the instrument are measured by the use of voltage dividers, whilst lower voltages must be suitably amplified.

Voltage transformers whilst extremely dependable are difficult to design to cover wide frequency bands and ratios, and in some circumstances their comparatively low impedance may be a source of error, as, for example, in determining current by the measurement of the voltage developed across a known standard resistor. The amplifier which is to be described gives a large number of fixed gains between 2 and 1 000 in the frequency band 30 c/s–30 kc/s and enables measurements of voltages lower than 60 volts to be made with accuracies approaching that of the instrument when used directly.

In order that it should be a satisfactory adjunct to the electrostatic voltmeters in use at the N.P.L. it should: (*a*) provide a number of accurately known gains in this frequency band and the ratio between successive gains should not exceed 1·6 : 1, and (*b*) be capable of producing an output voltage of 120 volts (r.m.s.) when connected to a load of 150–200 pF, which is the usual capacitance of the voltmeter and leads.

The method employed is that of using two dividers separated by two amplifiers with an additional amplifier at the input to act as a buffer stage giving the arrangement as shown in Fig. 1.

In order that the relative gains shall be determined solely by the dividers it is necessary to ensure that the amplifiers are sufficiently linear over their working voltage range and that their gains are independent of each other.

As the valves of the amplifier age the overall gain will change slowly with time and arrangements for calibration are therefore made. For ease of working and particularly to ensure confidence in the instrument it was considered essential to keep such operations to a minimum, and, accordingly, the initial target was to reduce the heating changes, as distinct from ageing effects, to about 1 part in 10^4 during a working period of 8 hours, and to an accuracy of 1 part in 10^4, to make the amplifier gains independent of frequency, output-voltage level and gain positions. A calibration check at one point would then give the performance at any combination of frequency, output-voltage level and gain.

(2) CONSIDERATIONS IN THE DESIGN OF THE AMPLIFIER UNITS

All three amplifier units are feedback amplifiers of a type commonly called the 'ring of three'. A detailed analysis of this circuit and other appropriate feedback theorems are given in Section 7. Before a detailed description of the complete amplifier is given, some relevant conclusions resulting from these considerations will be stated and discussed.

The circuit of the basic amplifier unit is shown in Fig. 2. The introduction of the overall feedback, β_3, across the resistor R_k, results in a subsidiary feedback loop in the circuit of V_1 involving the factor β_1.

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If $$A_1' = \frac{A_1}{1 + \beta_1 A_1(1 - \beta_3 A_3)}$$

where $\beta_1 = R_k/R_L$ and $\beta_3 = R_k/(R_f + R_k)$, then when $Z_s \to 0$ and $Z_g \to \infty$ the gain equation may be written very nearly as

$$A' = \frac{A_1' A_2 A_3}{1 + \frac{\mu_1 + 1}{\mu_1}\beta_3 A_1' A_2 A_3} \quad . \quad . \quad . \quad . \quad (1)$$

If the factor $(\mu_1 + 1)/\mu_1$ is neglected this expression is in the form of the general feedback equation with the exception that the denominator involves the 'harmonic' feedback[4] as distinct from the overall feedback $\beta_3 A_1 A_2 A_3$.

When $\beta_3 A_3 \to 1$, $A_1' \to A_1$ and the harmonic feedback becomes identical with the overall feedback. That is if $A_3 \to 1$ the subsidiary loop, β_1, becomes of minor importance when $\beta_3 \to 1$.

In the more general case when Z_s and Z_g are finite and not zero, it will be seen (Section 7.1.1) that they affect the gain, and one result is that the stability of the circuit becomes dependent upon the ratio Z_s/Z_g. It is clearly undesirable to restrict the source impedance as far as unit 1 is concerned, and there will be an appreciable, although known, variation in the source impedance of units 2 and 3. Thus the circuits must be carefully designed to minimize the effect of the term Z_s.

Eqn. (1) also shows that it is essential that μ_1 should be high, for, however large $A_1' A_2 A_3$, the circuit gain cannot be more constant or linear than $(\mu_1 + 1)/\mu_1$. Consideration of the formula for the input impedance (Section 7.1.2) also indicates another need for a high μ_1, for, if $\beta_3 A_1 A_2 A_3 \to \infty$, then from eqn. (4)

$$Z_{imax} = Z_s + (1 + \mu_1)Z_g \quad . \quad . \quad . \quad . \quad (2)$$

The grid–anode admittance of V_1 will result in additional parallel components across the input which have not been taken into account in the equations. However if, in order to obtain a high μ_1, V_1 is made a pentode, the effect of this admittance may be made small.

The input-impedance/frequency characteristic of the basic circuit (Section 7.1.2) is of particular importance with respect to units 2 and 3, the inputs of which shunt the precision variable dividers A and B.

The amount of overall feedback which may be applied in the circuit is related not only to the required stability margin but also to the flatness of the external gain/frequency characteristic and to F_e, the effective feedback, which is a measure of the effectiveness of the main feedback connection in reducing changes in amplifier gain.

Whilst F_e and the modulus of the harmonic feedback are the same at mid-frequencies it is shown (Section 7.2) that the value of F_e can be maintained over a wider bandwidth than that of the harmonic feedback. As this latter quantity is a measure of the distortion-reducing properties of the feedback loop, the distortion will rise towards the ends of the frequency band if F_e is taken as the criterion in assessing the frequency limits. This may not be serious provided that sufficient harmonic feedback can be applied to maintain the distortion below a certain maximum.

It is important that F_e shall not vary appreciably over the required bandwidth as otherwise the frequency/gain characteristic will change as the internal amplifier gain changes.

With respect to all the factors previously raised it is advantageous if one l.f. and one h.f. circuit are made of phase-advance structure. As a result of the cathode-follower action of V_3 the time-constants of V_2 and V_3 tend to be low and it is preferable to make the load of V_1 the dominant h.f. circuit.

(3) DETAILED DESIGN OF PRECISION AMPLIFIER

(3.1) Choice of Components

Whilst feedback reduces the effects of variations in value of components inside the feedback loop, the properties of these components are nevertheless of considerable importance. For instance, unless feedback factors of the order of 1 000 are applied, standard commercial components are not always adequate to ensure a sufficiently slow rate of change of performance. The most serious difficulty is that of reducing changes due to heating.

Changes in value of a component may arise as the result of a rise of ambient temperature inside the instrument case and also as a result of self-heating. As the total change due to heating is required not to exceed 1 part in 10^4, individual contributions must be of a very small order.

In the valve circuits variation may arise as the result of changes in any of the resistive elements. It is evident from the gain equation that the pentode will be more susceptible to load changes than a triode. Changes in anode current will arise owing to changes in the cathode bias resistor and screen bias resistor and cause gain changes owing to alteration of the valve parameters. Measurements of RdA/AdR for the various functions in an r.f. pentode gave values of 3/4 for the anode resistor, $-3/8$ for the cathode resistor and approximately 1/500 for the screen resistor under operating conditions similar to those to be used in the amplifier. For a medium-μ triode with R_L approximately $7r_a$, RdA/AdR was about 1/10 for the anode and $-1/10$ for the cathode resistor. As appreciable self-heating will arise only in the anode circuits, the load for the pentode will be the most vital consideration. In a particular instance in which the anode load dissipates 1·5 watts the resistor surface temperature is of the order of 100° C, and as the reduction in variation due to feedback is 70, the actual gain change will roughly equal the numerical value of temperature coefficient of the resistor.

The existence of positive and negative coefficients of RdA/AdR in the anode and cathode circuits suggests the possibility of compensation by choosing appropriate temperature coefficients. Resistance materials having positive and negative temperature coefficients could also be combined or used in different circuits. Such methods, however, are of doubtful value except perhaps where the correction required is small, since not only are matched temperature laws required but the rate of temperature rise must also be proportional. Unless the temperature rise can be sufficiently reduced the only satisfactory solution is the use where appropriate of resistance material of low temperature coefficient. In the example given in the previous paragraph a temperature coefficient not exceeding 10 parts in 10^6 per deg C is desirable.

The use of forced cooling was considered but rejected on several grounds, such as bulk and the effects of vibration, and other than promoting good natural convection cooling the performance is realized by suitable choice of components, being specially manufactured where necessary. Thus, whilst commercial h.s. carbon and vitreous wire-wound resistors to RCS/111 are found acceptable for screen and cathode circuits, respectively, those for the anode circuits are specially wound.

An additional source of variation of valve gain arises as a result of change of leakage resistance of coupling capacitors, which causes a change in valve bias and is again most serious when the valve is a pentode. The leakage resistance of a good-quality 0·5 μF paper capacitor was found to have fallen to about 3 000 megohms at 45° C, sufficient to cause a heating change of 0·5% in the gain of a following pentode where the grid leak was 1 megohm and the preceding potential 250 volts d.c. For this application plastic-film capacitors were found essential, whilst paper capacitors proved adequate when preceding triodes.

.eakage of the cathode by-pass capacitors will also cause iation of gain, and tantalum types have been used in the hope greater reliability.

'he main criterion in the choice of valves is that of linearity, rate of change of performance being reduced to an acceptable ue by moderate amounts of feedback. Where large voltage puts are required the triode is far more linear than the pen- e, although it will not produce as high a voltage before -off. It is also evident that, for a triode, the smaller is r_a greater is the voltage swing before cut-off. Consequently, ere the largest r.m.s. voltage is required, namely in unit 3, ver triodes are used in the output, and for the other units dium-μ voltage-amplifying triodes are used. As most distor- 1 occurs on the negative grid swing the operating voltage y be arranged with advantage according to the swing required. superiority of an r.f. pentode in the initial stage has already n discussed.

Short-term variations, arising in the feedback circuits or tage dividers, affect the amplifier gain directly, and these com- ients must be highly constant with both temperature and time. the long-term basis unchanging ratios of the dividers are the st vital consideration, since variation of amplifier gain will ur in any case. The design of these and other special resistors onsidered in the following Section.

(3.2) Special Resistors

The use of wire of high resistivity and low temperature coeffi- nt simplifies the design of high-frequency resistors, since for a en winding arrangement frequency errors tend to diminish as physical size of the resistor is decreased. The use of the aternary alloys, Karma and Evanohm, the properties of which ve been given elsewhere,[5, 6] help greatly in this respect.

All the resistors used are wound single layer on clear mica p of thickness 10 or 20 mils, according to wire diameter, and resistance wire is hard soldered to copper tails. The small- meter wire is protected from the sharp edges of the mica by ps of low-loss insulating material which are held in position the wire. Polythene is very suitable since it acts as a cushion, its use is limited to low temperatures. For the present her-temperature applications high-grade capacitor tissue is d. No varnish of any kind is applied to the main body of resistors, as it is found that, apart from the increased parallel acitance or possible strain effects, many varnishes are hygro- pic and cause the resistors to vary in a.c. resistance and time- istant. The use of clear mica is again essential, as small ourities were found to cause errors of several parts in 10^4 at kc/s on resistors of a few thousand ohms.

The special resistors fall into two classes. Those used as ode loads do not require to be accurately adjusted but merely retain their value and have a small temperature coefficient. ese are wound on 10-mil strip, $2\frac{3}{4}$ in $\times$ $\frac{1}{2}$ in, stiff copper minating leads being parallel to the short sides about $2\frac{1}{2}$ in rt so that the resistors fall directly into position on the tag p. Values of 50 and 300 kilohms are used and are wound ionally of 1-mil and 0·6-mil Karma wire, respectively, having nperature coefficients of less than 15 parts in 10^6 per deg C.

The remaining special resistors are the gain dividers and the stors in the feedback circuits.

The gain dividers at A and B in Fig. 1 are limited in value to otal resistance of 10 kilohms in order to avoid errors due to shunting effects at the upper frequency levels, and errors of capacitance due to switches, wiring and amplifier inputs. urther limit is set by the difficulty of designing a divider of ner value which at 30 kc/s will not be in error on account of own self- and earth impedances.

This latter difficulty also sets a limit to the maximum resistance

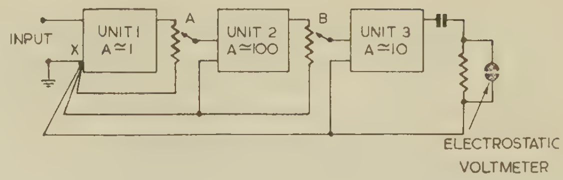

Fig. 1.—Arrangement of amplifier units.

of the feedback circuits, which are dividers formed by resistors such as R_f and R_k (Fig. 2), and where appropriate it is convenient to make one resistor combine two functions. Thus in unit 2 this resistor has additional tappings and serves as the gain divider B. In unit 1, where $\beta = 1$, the frequency response of the feedback circuit is not relevant as $R_f = 0$. However, parallel feeding of a gain divider at A is impracticable as the 30 c/s loss must be less than 1×10^{-4} and R_k is therefore made equal to 10 kilohms and tapped to serve this function.

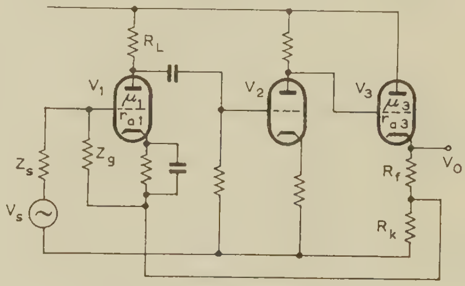

Fig. 2.—Basic amplifier circuit.

An unfortunate consequence of these arrangements is that as the resistors carry the direct current of the output valves their power dissipation is considerably higher than is usual in a precision resistor and is 6, 8 and 10 watts in units 1, 2 and 3, respectively.

The resistors are wound on strips of mica approximately 6 in $\times$ $1\frac{3}{4}$ in and No. 44 s.w.g. Evanohm is used throughout, except for some of the smaller values which exist in the resistor for unit 1. In order to provide reasonable facility for the adjustment of the lower values in this resistor, wire gauges up to No. 31 s.w.g. are used and fine adjustment is provided by constantan tails. With these lower values care has to be taken to ensure that the resistance contributed by the copper connec- tions does not unduly increase the temperature coefficient. The use of different gauges of wire increases the importance of the temperature coefficient of the main portion of this resistor. Temperature differentials will exist in all resistors and, coupled with the high power dissipation, make essential the use of wires with low temperature coefficients, especially in the case of the No. 44 s.w.g. Evanohm, which is of the order of 1 part in 10^6 per deg C.

Tappings on the various resistors are as follows:

Unit 1 tapped at $\frac{1}{5}$, $\frac{1}{10}$, $\frac{1}{50}$, $\frac{1}{100}$ and $\frac{1}{500}$.
Unit 2 tapped at $\frac{1}{100}$, $\frac{1}{3}$, $\frac{1}{2}$ and $\frac{3}{4}$.
Unit 3 tapped at $\frac{1}{10}$.

The initial tapping on unit 2 and the tapping on unit 3 are the feedback connections and are not accurately adjusted; the adjustment of the remaining tappings is better than ± 3 parts in 10^5.

(3.3) Power Supply

A stabilized h.t. power supply of 550 volts feeds all three units, which are individually decoupled for high frequencies only.

Decoupling at low frequencies requires components of large value, and might also interfere with the low-frequency stability and response of the individual units. A power supply with an impedance low enough to avoid mutual coupling at low frequencies is therefore used. To reduce the possibility of common paths at high frequencies the h.t. and earthed rails are run separately to each unit from common input points.

Since the instrument is to be used with inputs at 50 c/s it is necessary to reduce the mains-frequency components to a very low level to minimize the magnitude of beats. As there is no filtering subsequent to the h.t. supply it is necessary to reduce the hum level of this supply to about 100 μV for the hum level in the output to be 90 dB down. In order to avoid hum arising from valve heaters, d.c. heating is employed using a 300-volt commercial stabilized supply. This supply also feeds the amplifying section of the 550-volt h.t. supply, with the exception of the series valves. The stabilization of the supply to the heaters is also necessary, in addition to the stabilization of the h.t. supply, to prevent variation in amplifier gain with variation of supply voltage.

The circuit employed combines a negative-feedback loop from the stabilizer output and a forward path from the input. Two type 83A1 reference tubes are used in series and enable about one-third of the output to be sampled. This voltage is amplified by a pentode and passed to one grid of a long-tailed triode pair. The other grid of the long-tailed pair is used to connect in a variable forward-feedback path, and the voltage from the appropriate anode is connected to the grid of the series valve. As a result of forward feedback, the impedance in the input side of the stabilizer can be virtually eliminated and the impedance at low frequencies kept low down to zero frequency. The same setting also very nearly cancels hum due to the input circuit and also increases the stabilization against mains changes. The remaining hum was reduced by some elementary electromagnetic screening, valves being the main offenders in this respect. The residue above noise level is largely due to the series valve, which is heated by a mains frequency voltage. As the valves in the stabilizer are heated from a stabilized d.c. supply there is no necessity to compensate for heater variations.

The overall stabilization is 1 000 when the forward path is zero. When the forward path is correctly adjusted stabilization is increased by about an order; individual hum components are not greater than 20 μV, and the output impedance from d.c. to 30 kc/s supply at the input terminals to the amplifier is not greater than 0·1 ohm. The change of output over an 8-hour period is about 0·1 %, which is sufficient to change the amplifier gain by about 1 part in 10^5.

(3.4) Amplifier Units

(3.4.1) Unit 1 Gain $\simeq$ 1.

The main problem in the initial unit is that of ensuring sufficient linearity over the full range of voltage and frequency. As 100 % feedback is employed a high loop gain is easily obtained. This is reduced to about 57 dB by the application of a subsidiary loop over the second stage, which not only eases the stability considerations but retains the advantage of the original loop gain as far as this stage is concerned.

Over the input range of 60 mV to 60 volts the errors due to non-linearity are a few parts in 10^5 at middle frequencies and rise to rather less than 1 part in 10^4 at 30 kc/s. This increase is probably entirely due to the decrease of the modulus of the loop gain, but there may also be some contribution from the initial stage, which is working into capacitive loads at higher frequencies.

The components in the input circuit are screened to the output terminal in order to reduce the input capacitance. As th increases the grid–cathode capacitance, C_g, it somewhat adverse affects the input resistance at higher frequencies. The input lea is low-capacitance screened cable terminated at the measuri end by a probe. The screen may be connected either to eart giving a coaxial input system, or to the cathode, in which ca a separate earth lead must be used, giving input capacitances 15 pF and 2–3 pF, respectively. In the latter case the inp resistance is approximately −50 megohms at 30 kc/s. In cases the mid-frequency input resistance is about 70 megohms.

The valves used are one r.f. pentode, type UF80, and two r double triodes, type UCC85, the output or cathode-follow stage comprising three triodes in parallel.

(3.4.2) Unit 2 Gain $\simeq$ 100.

As unit 2 employs only 1 % feedback it is not possible realize a high loop gain. The output voltage varies betwee and 36 and sufficient linearity is obtained with a harmonic lo gain of 34 dB.

The input resistance varies from 70 megohms at low fr quencies to about −80 megohms at 30 kc/s and the inp capacitance is about 5 pF.

The valves are as used in unit 1.

(3.4.3) Unit 3 Gain $\simeq$ 10.

As the overall feedback is 10 % R_k (Fig. 2) is now sufficie high to cause appreciable subsidiary feedback unless the resistance of V_1 is also high. Further, as power triodes are us in the output stage where voltages up to 120 volts r.m.s. a required, the harmonic loop gain is again reduced. To ob a high gain from the initial stage a triode cathode-follower h been added enabling a value of 300 kilohms to be used for t anode load of the initial pentode. In this way a harmonic lo gain of approximately 38 dB is obtained.

The output voltage at the cathode of the final valve is appl to the electrostatic voltmeter terminals by a *CR* circuit as sho in Fig. 1. As the voltmeter is an r.m.s. indicating instrumen is necessary to avoid any appreciable direct voltage appearing the voltmeter terminals, and at the same time it is required transmit with negligible error frequencies down to 30 c/s. Th whilst the *CR* product should be high on this count, a large C product may cause errors due to d.c. leakage of the capacito A compromise is reached using values of 2 μF and 220 kilohm Consequent upon this arrangement the advantage of the very lo output impedance is lost and the amplifier output at the vo meter terminals is adversely affected by parallel loading. A apparent loss of gain of 1 part in 10^4 occurs at all frequencies f each 200 pF of additional loading.

Provision for adjustment of gain is made by circuits in paral with the tapping section of the feedback resistor so that t overall gain may be adjusted to exactly 1 000.

The input resistance of this unit varies from abo 140 megohms at low frequencies to 100 megohms at 30 kc and the input capacitance is about 5 pF.

The valves used are one r.f. triode-pentode, type UCF80, a two power triodes, type UL84.

(3.5) Constructional Details

The power supply and the amplifier form separate un mounted on a common trolley, the power supply near grou level and the amplifier at bench level. Sufficient connecting le is provided to make it possible to remove the amplifier and pla it near to the work.

The three amplifier units are separately constructed and co tained in a single case which has three screened vertical sectio Adequate ventilation is encouraged by mounting the valv

ermost and allowing free upward flow through each section. most vital components, namely dividers and feedback stors, are placed at the base in electrostatically screened, ilated boxes. The remaining amplifier components are inted vertically above the boxes on skeleton tag-strip mblies and below the valve line. Ceramic insulated tag is used for components in the input grid circuits, and p.t.f.e. e bases are used throughout.

igh-grade insulating materials are also required in the power ply as leakages may cause an increase of hum voltage.

(3.6) Calibrating Circuit

alibration is effected by deriving a voltage from a divider ing the inverse ratio to the amplifier gain. The divider input amplifier output are displayed on two electrostatic voltmeters will give equal deflection when the gain is nominal. When plies are sufficiently steady, the voltages may be displayed rnately on the same voltmeter.

he complete range of gains is covered by three tapped dividers hat a calibration may be made on any range and at any uency in the band. Whilst a calibration on a single range ild be sufficient to determine all the other ranges, it is never-ess essential to provide facilities for making completely pendent checks to confirm that the instrument is continuing vork satisfactorily.

he construction of the calibration dividers is similar to that he comparable dividers in the amplifier itself.

is essential that the earth connection shall be made only at junction of the earthed input lead and the test circuit. If the rument end of the earth lead, point X in Fig. 1, is connected arth—a very desirable arrangement from several points of —earth current from the test circuit will be drawn down earthed input lead and cause errors dependent upon fre-ncy, instrument range and the nature of the test circuit.

(3.7) Performance

fter adjustment to nominal gain on the ×1 000 range at volts output and 1 kc/s frequency:

	Parts in 10^4
ain on all ranges at 1 kc/s	± 1
requency response all ranges 30c/s–10kc/s	± 1
requency response all ranges 30c/s–30kc/s	± 2
elf-heating change per 8-hour period (after initial 5 min warming up)	$\not> \pm 1$
ain constancy per 100 hours intermittent use	$\not> -2$
hange in gain between 60 volts and 120 volts output	$\not> \pm 1$
hanges due to mains variation	negligible

hese values were not exceeded during 500 hours' use spread a period of 3 years.

(4) CONCLUSIONS

has been shown that a multi-gain amplifier having highly tant performance can be constructed using simple circuits with moderate amounts of feedback, provided that due care aken in the choice of components. Without the use of lern resistance materials it is unlikely that with the present gn equal performance could have been obtained with respect oth self-heating and frequency response.

his valve amplifier has been proved to be a satisfactory means creasing the sensitivity of a precision electrostatic voltmeter, he type used at the N.P.L., without greatly reducing its racy or frequency range.

s the intrinsic value of both the instrument and the work for which it will be employed is high, complete and fairly frequent valve replacement will be justified. Thus with replacement after not more than 1 000 hours' use very little deterioration of properties is expected.

(5) ACKNOWLEDGMENTS

The work described above has been carried out as part of the research programme of the National Physical Laboratory, and this paper is published by permission of the Director of the Laboratory.

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(7) APPENDICES

(7.1) Circuit Analysis

The equivalent circuit for determining the gain and input impedance is shown in Fig. 3: no circuit is shown for the derivation of the input impedance.

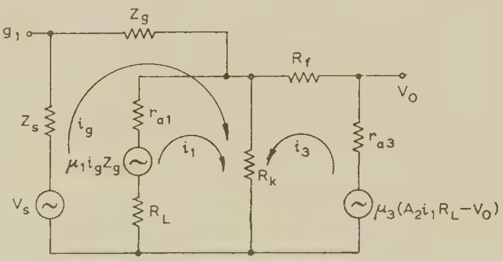

Fig. 3.—Amplifier equivalent circuit.

(7.1.1) Amplifier Gain.

An exact solution is cumbersome, but with very small error in a practical case,

$$A' \simeq \frac{A_1 A_2 A_3}{1 + P + Q + R + S\beta_3 A_1 A_2 A_3} \quad . \quad . \quad . \quad (3)$$

where
$$A_1 = \frac{\mu_1 R_L}{r_{a1} + R_L + R_k} \qquad \beta_1 = \frac{R_k}{R_L}$$

$$A_3 = \frac{\mu_3 (R_k + R_f)}{(\mu_3 + 1)(R_k + R_f) + r_{a3}} \qquad \beta_3 = \frac{R_k}{R_f + R_k}$$

$$P = \frac{Z_s}{Z_g}\left(1 + \frac{\beta_3 A_1 A_2 A_3}{\mu_1}\right)$$

$$Q = \frac{(1 - \beta_1 A_3)R_k}{Z_g}$$

$$R = \beta_1 A_1 (1 - \beta_3 A_3)$$

$$S = \frac{\mu_1 + 1}{\mu_1}$$

The significance of the terms R and S has been discussed in Section 2.

The term P shows the influence of the source impedance Z_s. If the amplifier has an input capacitance C_s this may be included in the source Z_s. At low frequencies the factor Z_s/Z_g becomes $(R_s + 1/j\omega C_s)/R_g$, which produces an imaginary term due to the time-constant C_sR_g. This affects stability in a manner roughly equivalent to the addition of a further time-constant in the amplifier gain $A_1A_2A_3$. Increase of R_s reduces the effect of this and thereby improves stability.

At high frequencies Z_s/Z_g becomes $j\omega C_g Z_s$, where C_g is the parallel capacitive component of Z_g, and not only adversely affects stability but varies directly with Z_s. Thus an amplifier which is stable with the input short-circuited may well oscillate when closed by some finite impedance. In circumstances where Z_s is varied the effect of this term must be minimized, which can best be effected by increasing the major high-frequency time-constant in the amplifier as much as possible.

(7.1.2) **Input and Output Impedances.**

The input and output impedances are given very nearly by

$$Z_i = \frac{v_s}{i_g} \simeq Z_g \frac{A_1A_2A_3}{A'} \frac{1}{1 + \beta_3 A_1 A_2 A_3/\mu_1} \quad . \quad . \quad (4)$$

$$Z_o \simeq \frac{r_{a3}}{\mu_3 A_3 (1 + \beta_3 A_1 A_2)} \quad . \quad . \quad . \quad . \quad (5)$$

The variation of input impedance with increasing frequency may be readily obtained by expressing the gains A_1A_2 and A_3 in eqn. (4) in terms of the time-constants of the circuits. As a consequence of the appearance of negative real quantities on multiplication, the resistive component of the input impedance rises to infinity and then becomes decreasingly negative. Partial correction for this effect is obtained by connecting an RC branch in parallel with the input.

If the time-constants in the gain circuits are well staggered the frequency at which the uncompensated value of R_i is infinite tends to be independent of βA_m and almost directly dependent on the product of the major gain circuit and the grid circuit time-constants. Thus, the minimum acceptable value of R_i having been decided, the maximum useful frequency is limited by the extent to which C_g can be reduced.

(7.2) **Effective Feedback**

The extent by which variations of gain inside the feedback loop are reduced by the application of feedback is called the effective feedback and denoted by F_e.

Thus $$F_e = \frac{dA/A}{|dA'/A'|} = \frac{dA_m/A_m}{|dA'/A'|} \quad . \quad . \quad . \quad . \quad (6)$$

In the general feedback equation we may write the internal gain, A, in terms of A_m, the mid-frequency gain, and the circ[uit] time-constants. Thus, for a system having three time-consta[nts] τ, $n_2\tau$ and $n_3\tau$, in which β is real and negative, we have, at h[igh] frequencies,

$$A' = \frac{A_m}{(1 + j\omega\tau)(1 + j\omega n_2\tau)(1 + j\omega n_3\tau) + \beta A_m}$$

After expanding and separating real and imaginary terms, [an] equation for A' may be written, which by differentiation gi[ves] $|dA'|/dA_m$.

From this it follows that

$$F_e = 1 + K\beta A_m, \quad . \quad . \quad . \quad . \quad .$$

where

$$K = \frac{1 - \omega^2\tau^2(n_2 + n_3 + n_2n_3) + \beta A_m}{1 + \beta A_m + \omega^2\tau^2[(1 + n_2^2 + n_3^2) - \beta A_m(n_2 + n_3 + n_2n_3)] + \omega^4\tau^4(n_2^2 + n_3^2 + n_2^2n_3^2) + \omega^6\tau^6 n_2^2 n_3^2} \quad . \quad . \quad .$$

The magnitude of K, which is unity at mid-frequencies, [will] vary as the frequency rises in a manner determined by [the] coefficients of the terms in $\omega^2\tau^2$. Under some conditions it [will] simply fall steadily from the mid-frequency value, whilst un[der] others it will rise to infinity and then become negative, as sh[own] in Fig. 4.

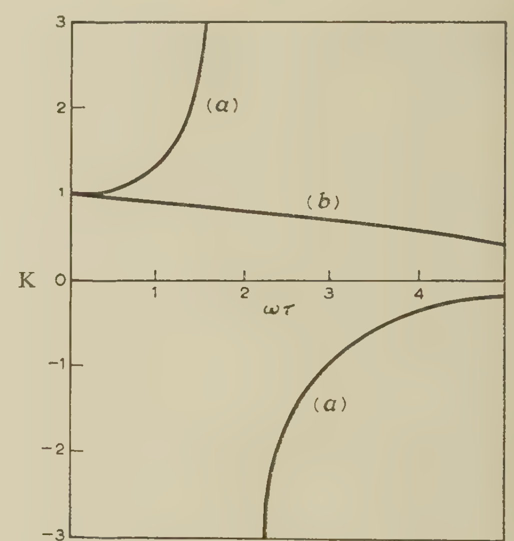

Fig. 4.—Variation of K with frequency.

(a) $\beta A_m = 70$, $n_2 = \frac{1}{40}$, $n_3 = \frac{1}{4000}$

(b) $\beta A_m = 25$, $n_2 = n_3 = \frac{1}{50}$

Point $\omega\tau = 1$ corresponds to the frequency at which $|\beta A|$ has fallen by 3 dB

If the dominant time-constant in the preceding analysis [is] replaced by a phase-advance network, K may be obtained [by] converting the network into an equivalent parallel RC netw[ork] for each specific frequency. With the notation given in Fig[.] and putting $R_1C_2 = \tau$, the impedance between the terminals

$$Z = \frac{a_\omega R_1}{1 + j\omega\tau_\omega} \quad . \quad . \quad . \quad . \quad . \quad ($$

where $$a_\omega = 1 - \frac{\omega^2\tau^2 y}{1 + \omega^2\tau^2 y(1 + y)} \quad . \quad . \quad . \quad ($$

and $$\tau_\omega = \tau(1 + x)\left\{1 - \frac{(1 + x + y)\omega^2\tau^2 y}{(1 + x)[1 + \omega^2\tau^2 y(1 + y)]}\right\} \quad . \quad ($$

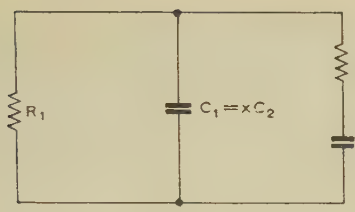

Fig. 5.—Phase-advance network.

If $y \ll 1$, which is the most usual condition,

$$\tau_\omega \simeq \tau(1+x)a_\omega \quad . \quad . \quad . \quad . \quad . \quad . \quad (13)$$

If the remaining time-constants in the circuit are expressed in
rms of τ, noting from eqn. (10) that the gain is multiplied
a_ω,

(7.3) Frequency Response

In the present amplifier a flat frequency/gain characteristic is required over the working band to within about 1 part in 10^4, and in order to measure the r.m.s. value of a voltage irrespective of waveform it is also necessary to maintain a reasonably flat response well outside the working band. This second condition is likely to follow to some degree if the first condition is realized.

The problems of design are appreciably different at the two ends of the frequency band. The gain equation in terms of seen components may be simpler at the high-frequency end; however errors due to unseen components may arise in this case, and are unlikely at the low-frequency end. Small corrections of the response may be easily made at high frequencies, whereas at low frequencies they are likely to require physically large components.

The optimum design does not necessarily coincide with the

$$= \frac{a_\omega[1-\omega^2\tau_\omega^2(n_2+n_3)-\omega^2\tau^2 n_2 n_3+a_\omega\beta A_m]}{1+a_\omega\beta A_m+\omega^2\tau_\omega^2[1-a_\omega\beta A_m(n_2+n_3)]+\omega^2\tau^2(n_2^2+n_3^2-n_2n_3a_\omega\beta A_m)+\omega^4\tau_\omega^4(n_2^2+n_3^2)+\omega^4\tau^4n_2^2n_3^2+\omega^6\tau^6n_2^2n_3^2} \quad . \quad (14)$$

For given values of τ, n_2 and n_3, K, as given by eqn. (14),
ll fall more rapidly as the frequency increases than does the
evious value given by eqn. (9). Since, however, the use of the
ase-advance network will reduce the value of τ, i.e. the
minant time-constant required for a given stability margin,
e actual fall in two comparable cases may not be very dissimilar.
s a consequence of the reduced value of τ the modulus of the
op gain falls less rapidly with frequency and will thus reduce
e variation in distortion over a given frequency range.

In the unity-gain amplifier described in Section 3.4.1, the
llowing conditions apply; $\beta A_m = 700$, $x = 1/20$, $y = 1/80$,
$= 1/100$, $n_3 = 1/20\,000$, and K varies as shown in Fig. 6.
ne value of $\omega\tau = 1$ occurs at 7 500 c/s and corresponds to a
ss of 3 dB in $|\beta A|$. At 30 kc/s the reduction in K is negligible
hilst $|\beta A|$ has now fallen by about 12 dB.

maximally flat condition, since the actual errors at the required limits of frequency may exceed those permissible. In practice it is feasible to allow for some correction at the high-frequency end of the band, so that a design avoiding h.f. response peaks is simplified.

For the circuit discussed in the previous Section having two simple time-constants and a phase-advance network, we may write

$$\frac{A_m'}{A_h'} = \frac{\frac{1}{a_\omega}(1+j\omega\tau_\omega)(1+j\omega\tau n_2)(1+j\omega\tau n_3)+\beta A_m}{1+\beta A_m} \quad . \quad (15)$$

where a_ω and τ_ω are as before.
From which

$$\left|\frac{A_m'}{A_h'}\right|^2 = 1 + \frac{\frac{1}{a_\omega^2}(1+\omega^2\tau_\omega^2)(1+\omega^2\tau^2n_2^2)(1+\omega^2\tau^2n_3^2) - 2\frac{\beta A_m}{a_\omega}[\omega^2\tau\tau_\omega(n_2+n_3)+\omega^2\tau^2n_2n_3-1+a_\omega]-1}{(1+\beta A_m)^2} \quad . \quad . \quad (16)$$

The value of F_e for low frequencies for an equivalent circuit
ay be readily obtained. In practice, the low-frequency circuit
much more complicated owing to the existence of screen and
thode by-passing. Although F_e may still fall less rapidly

If $|A_m'/A_h'|$ is close to unity, the deviation from unity, $\Delta|A_m'/A_h'|$, is very nearly one-half the right-hand term.

When $y \ll 1$, by eqn. (13) we may write, after separating positive and negative terms,

$$\left|\frac{A_m'}{A_h'}\right| \simeq \frac{\left[\frac{1}{a_\omega^2}+\omega^2\tau^2(1+x^2)\right](1+\omega^2\tau^2n_2^2)(1+\omega^2\tau^2n_3^2)+2\frac{\beta A_m(1-a_\omega)}{a_\omega} - 2\beta A_m\omega^2\tau^2(1+x)(n_2+n_3)+\frac{\omega^2\tau^2n_2n_3}{a_\omega}-1}{2(1+\beta A_m)^2} \quad . \quad (17)$$

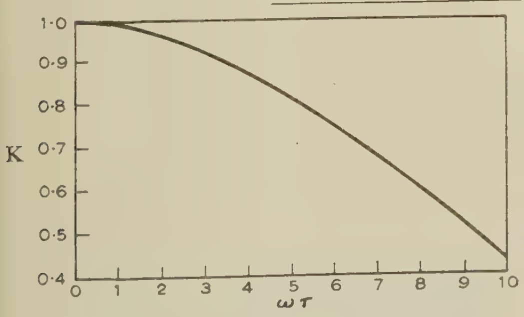

Fig. 6.—Variation of K for amplifier unit 1.

an $|\beta A|$ for decreasing frequency, this relationship is involved
nly to a small extent in the amplifiers described, where it has
een necessary to maintain $|\beta A|$ at the lower frequencies to
btain the necessary frequency response.

If the phase-advance network is converted to a single time-constant by making $y = 0$, then $a_\omega = 1$ and $\tau_\omega = (1 + x)$. The positive part of the numerator of eqn. (18) is acting so as to reduce the high-frequency gain and contains the principal contribution due to the phase-advance network. Whilst the principal function of the phase-advance network is that of shaping the phase-amplitude response at frequencies remote from the working band to achieve stability and minimize response peaks, the contributions inside the working band may be appreciable. If the resulting characteristic shows a fall with frequency, fairly accurate correction may be made by a series RC network across the feedback tapping if the original error does not exceed 0·1%. If $\beta \to 1$, however, this method is not applicable.

Where the feedback resistor is used additionally as a tapped divider across the output to provide a number of different gains, only minor correction to the feedback circuits may be made as

the frequency characteristics of the various gain positions must not be impaired. For a number of cascaded amplifiers, however, individual correction need not necessarily be made. It seems likely that with a more complicated phase-advance structure the gain/frequency characteristic could be improved, for example by making y a frequency variable.

Measurement of the frequency/gain characteristics of the amplifiers described in Section 3.4 shows that in the worst case there is a fall of 0·1% at the highest frequency without correction. The calculated values agree to within a few parts in 10^4 with the measured values. As the effect of stray capacitances cannot be accounted for the agreement is surprisingly good.

A similar analysis may be made at low frequencies. An exact solution including the effects of cathode and screen by-passes in addition to inter-stage couplings is, however, considerably involved. For a minimum frequency of 30 c/s it is practicable to use couplings sufficiently large for the error due to them at this frequency to be significantly less than that due to screen and cathode by-passes.

In the absence of screen and cathode by-passes the circuit reduces to two time-constants, τ and $n_2\tau$, τ being the dominant and thus smaller time-constant. By an analysis similar to the mid-frequency case,

$$\Delta\left|\frac{A_m'}{A_l'}\right| = \frac{\left(1 + \frac{1}{\omega^2\tau^2}\right)\left(1 + \frac{1}{\omega^2\tau^2 n_2^2}\right) - 2\frac{(1 + \beta A_m)}{\omega^2\tau^2 n_2} - 1}{2(1 + \beta A_m)^2} \quad . \ (18)$$

As an example, suppose $\beta A = 50$, $\tau = 0 \cdot 1$ sec and $n_2 =$ then at 30 c/s the error is roughly 1 part in 10^5. The princip design problem is then to provide adequate by-passing in scree and cathode circuits. As the relation between phase and amp tude depends upon the valve and circuit parameters a gener design procedure is difficult. If the gain of the circuit excludi the coupling capacitors is calculated from the standard formula and expressed at the lowest frequency as

$$A = A_m\frac{a + jb}{c + jd} \quad . \quad . \quad . \quad . \quad . \quad (1$$

then we can write

$$\left|\frac{A_m'}{A_l'}\right|^2 = 1 - \frac{\frac{c^2 + d^2}{a^2 + b^2} - 1 + 2\beta A_m\left(\frac{ac + bd}{a^2 + b^2} - 1\right)}{(1 + \beta A_m)^2} \quad . \ (2$$

When $|A_m'/A_l|$ is near unity the deviation is again one-half t right-hand term. The relative importance of the various ter in the gain equation will be evident from inspection, and appropriate choice of time-constants for the screen and cath circuit the coefficients a, b, c and d may be adjusted so that t deviation from unity of $|A_m'/A_l|$ is sufficiently small. The add tional loss due to the dominant coupling time-constant circu may then be introduced in the gain equation to check if further adjustment of values is desirable. Stability is improv by making the dominant coupling a phase-advance netwo without materially affecting the gain in the working band.

[The discussion on the above paper will be found on page 337.]

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THE DESIGN AND PERFORMANCE OF HIGH-PRECISION AUDIO-FREQUENCY CURRENT TRANSFORMERS

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SUMMARY

Current transformers of existing design are found to have large high-frequency errors. The causes of these errors are examined and it is shown that they are due to the high values of leakage inductance, self-capacitance and concentrated inter-winding capacitance that result from the conventional multi-layer form of construction. Greatly improved performance can be obtained if the windings are arranged as uniformly-distributed single layers with a suitable thickness of insulation between them. Approximate formulae for calculating the various constants and errors of a transformer having single-layer windings are given, together with a design procedure for any specified limit of error over a wide band of frequencies. The practical aspects of the design are also discussed.

Multi-ratio transformers in which all the ratios have virtually identical errors over a band of frequencies can be obtained by subdividing single-layer primary windings into sections, all of which are used by means of series-parallel connections for each ratio.

Details are given of the design and performance of precision single-layer multi-ratio transformers rated from 5/5 to 400/5 amp. Their errors do not exceed 5 parts in 10^5 in ratio and $0\cdot3'$ in phase over the frequency range 400 c/s–10 kc/s and are less than 2 parts in 10^4 and over the range 50 c/s–30 kc/s.

LIST OF PRINCIPAL SYMBOLS

I_p = Current at primary terminals.
I_s = Current at secondary terminals.
I_p/I_s = True ratio.
N_p = Number of primary turns.
N_s = Number of secondary turns.
$T = N_s/N_p$.
R_p = A.C. resistance of primary winding.
R_s = A.C. resistance of secondary winding.
R_b = A.C. resistance of secondary burden.
L_p = Leakage inductance of primary winding.
L_s = Leakage inductance of secondary winding.
L_b = Inductance of secondary burden.
$Z_p = R_p + j\omega L_p$.
$Z_s = R_s + j\omega L_s$.
$Z_b = R_b + j\omega L_b$.
$Z = Z_s + T^2 Z_p$ = Total impedance referred to secondary winding.
Z_c = Effective impedance for core magnetizing and loss currents.
C_p = Effective lumped self-capacitance of the primary winding.
C_s = Effective lumped self-capacitance of the secondary winding.
C_y = Total capacitance between the windings.
Y_p, Y_s, Y = Admittances of C_p, C_s and C_y respectively, neglecting loss components.
ϵ = Permittivity of the interwinding insulation.
$\omega = 2\pi f$, where f is the frequency in cycles per second.

The paper is an official communication from the National Physical Laboratory.

(1) INTRODUCTION

During the past few years there has been an increasing demand for the accurate measurement of alternating currents up to a few hundreds of amperes in the audio-frequency range of 400 c/s–10 kc/s. For currents greater than about 5 amp, current transformers have been used and the Laboratory has been called upon to test these instruments, usually up to $2\cdot4$ kc/s but occasionally at higher frequencies. The majority of the transformers were toroidally wound and purported to be of precision grade yet were found to be wholly unsuitable for accurate measurement in the audio-frequency band.

In most cases it was found that as the test frequency was raised

(*a*) The secondary current became greater than its nominal value with a lagging phase angle, the errors increasing approximately as the square of the frequency.

(*b*) The errors became increasingly dependent upon the circuit conditions, being different for each of the four possible positions of a connecting link between the primary and secondary circuits.

(*c*) The winding resistance increased by a factor of two or more, developing unduly high internal temperatures.

(*d*) The primary voltage required to supply the transformer increased to a prohibitively high value.

(*e*) In the case of tapped-primary multi-ratio instruments, the errors for the various ratios became increasingly different.

In the following Sections the causes of the errors and defects which occur in transformers of existing conventional design are examined and means are indicated whereby they may be reduced. The design criteria for standard transformers capable of operating over a wide range of frequencies with very high accuracy are discussed, and finally the details and performance of a range of instruments that have recently been constructed at the Laboratory are given.

(2) THE CAUSES OF HIGH-FREQUENCY ERROR

The well-known text-book theory of current transformers takes into account only the effects of the core exciting currents, the impedances of the windings and the secondary burden, and therefore gives no indication of the causes of the phenomena which have been listed in Section 1. In an exhaustive analysis, however, Arnold[1] has considered the effect of capacitance on the performance of multi-layer toroidally-wound transformers. Assuming that the capacitance currents are small compared with the main currents, he has developed a series of equations for various distributions of the inter-winding capacitances, from which the transformer errors may be deduced. In view of the number and complexity of the equations it is not proposed to reproduce them here, and reference should be made to the original paper for the details. They are, however, of the general form

$$\frac{\text{True ratio}}{\text{Turns ratio}} = 1 + \frac{Z_b + Z_s}{T^2 Z_c} + \frac{Y_p}{T^2}(Z + Z_b) + Y_s Z_b + \frac{Y}{T}(Z + Z_b)\left(\frac{k_1}{T} \pm k_2\right) + \frac{Z_b Y}{T}(k_3 \pm T k_4) \quad . \quad (1)$$

where k_1, k_2, k_3 and k_4 are constants derived from expressions which involve the distribution of the inter-winding admittance Y.

The authors of the present paper have carried out experimental work on several toroidal transformers with multi-layer windings and have found that the ratio and phase-angle errors calculated from the appropriate equations were in very close agreement with the measured values. It was further found that, with increasing frequency, the bulk of the overall error was accounted for by the effects of leakage inductance and winding resistance combined with the value and distribution of the inter-winding capacitance, i.e. by the term in eqn. (1) which involves the product YZ. The overall values obtained in one typical case are given in Table 1. Each winding of this transformer had

Table 1

Frequency	Measured values		Calculated values	
	True ratio / Turns ratio	Phase angle	True ratio / Turns ratio	Phase angle
With terminals M and Ⓛ linked				
c/s		minutes		minutes
400	0·9997	−0·1	0·9997	−0·2
1 000	0·9982	−1·1	0·9982	−1·0
2 000	0·9928	−4·0	0·9927	−3·8
With terminals L and Ⓜ linked				
400	0·9999	0·0	1·0000	0·0
1 000	0·9997	+0·1	0·9998	−0·1
2 000	0·9989	−0·4	0·9991	−0·4

1 000 turns uniformly distributed around the core in five layers. The total leakage inductance was 5·5 mH, the inter-winding capacitance was 5 000 pF, mainly concentrated between the L, and Ⓜ terminals, and the resistance of the windings at 2 kc/s was $3\frac{1}{2}$ times the d.c. value. Attention is drawn to the magnitude of the errors and their dependence upon the circuit conditions: such large errors are a direct consequence of the effects that result from the use of the conventional multi-layer form of construction.

(3) DESIGN CONSIDERATIONS FOR WIDE-FREQUENCY-BAND OPERATION

It was considered necessary for the Laboratory to have standard current transformers, ratios 5/5 to 400/5, whose errors did not exceed 1×10^{-4} in magnitude or phase for any method of circuit connection throughout the frequency band 400 c/s–10 kc/s and, if possible, over an even wider band. In order that transformers complying with this specification may be constructed it is first necessary to find a design that will give a value for the product of leakage inductance and inter-winding capacitance several orders of magnitude lower than that given by existing designs. It is also desirable that the new design should result in a favourable distribution of the capacitances. It will be shown that these requirements can be realized by adopting a single-layer method of winding, and the following Sections deal with the theoretical and practical aspects of such transformers.

Examination of Arnold's equations shows that the optimum performance of a transformer is obtained if the leakage inductances, inter-winding capacitances and resistances are uniformly distributed. In this case eqn. (1) becomes

$$\frac{\text{True ratio}}{\text{Turns ratio}} = 1 + \frac{Z_b + Z_s}{T^2 Z_c} + \frac{Y_p}{T^2}(Z + Z_b) + Y_s Z_b + \frac{Y}{12T^2}[Z(1 - T) + Z_b(1 - T)^2] \quad . \quad (2)$$

Eqn. (2) applies when there is no direct connection betwe the primary and secondary circuits. When these are joined a low-impedance link the last term in eqn. (2) must be modified

$$\frac{Y}{12T^2}[2Z(2 + T) + 4Z_b(1 + T + T^2)]$$

when the link connects M and Ⓛ or L and Ⓜ

or
$$\frac{Y}{12T^2}[4Z(1 - T) + 4Z_b(1 - T)^2]$$

when the link connects M and Ⓜ or L and Ⓛ.

The necessary condition of uniform distribution can be reali in practice by winding the primary and secondary turns each i single layer around the core.

In the present application we are only concerned with st down or unity-ratio transformers. The maximum value of a form of the important last term in eqn. (2) is then $Y(Z/2 + Z$ this occurring when the turns ratio, T, has its minimum value unity. Since standard transformers are invariably used in t laboratory with non-inductive burdens, the maximum error a the maximum change in error that can occur due to this te may be written as $-\frac{1}{2}\omega^2 C_y(L_p + L_s) + j\omega C_y(R_p + R_s + 2R$ It will be convenient to consider the two components of t error separately.

(3.1) Leakage Inductance in Toroidal Transformers

In the following treatment it is assumed that the second turns are wound next to the core and that the conductors winding are so close together that the leakage due to the around the individual wires may be neglected.

(3.1.1) Primary Winding.

The primary leakage inductance of a toroidal transfor having multi-layer windings may be considered to be due to th groups of flux paths:

(*a*) Those which link all the primary turns but only part of t secondary.
(*b*) Those which link all the primary but none of the seconda
(*c*) Those which link only a part of the primary turns.

In such transformers the depth of the windings is usually lar compared with the thickness of the inter-winding space, a while this condition holds, variation in the dimension of t space will have little effect on the leakage inductance. As t winding depth is reduced the inductance falls steadily and t inter-winding space assumes an increasing importance. T limit is reached when the depth is reduced to that of a sing layer.

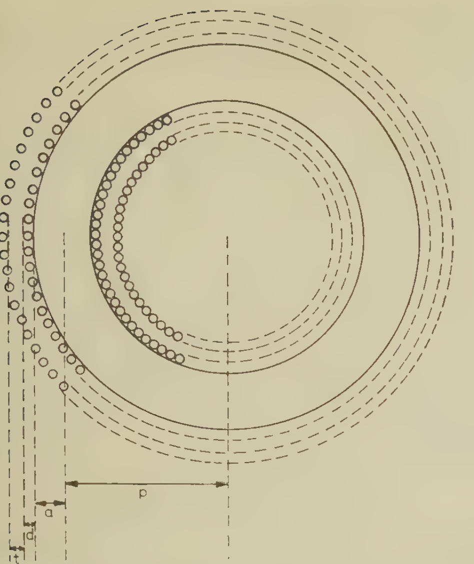
Fig. 1 shows the cross-section of a single-layer unity-ra transformer having all sections circular. If it is assumed th all the leakage flux is located in the annular space defined by t axes of the conductors, then it can be shown that the prima leakage inductance is given by

$$L_p = \frac{2\pi N_p^2}{p}(2a + 2d + t)(t + d) \times 10^{-9} \text{ henry} \quad . \quad ($$

In practice, the section of the core will be rectangular rath than circular, but eqn. (3) may still be used with reasonab accuracy if $1/\sqrt{\pi}$ times the square root of the core area is take for the value of the dimension a. The manner in which t inductance is dependent upon the thickness of the inter-windi space, t, will clearly be noted.

(3.1.2) Secondary Winding.

It is generally conceded that the secondary leakage inductan of a toroidal transformer is small if both windings are uniform



ig. 1.—Dimensions of transformer with single-layer windings and all sections circular.

tributed around the core. In the case of single-layer-wound truments, analysis of the errors has shown that the secondary kage could not have exceeded 5% of the total. For such nsformers the secondary leakage can therefore be neglected.

(3.2) Inter-winding Capacitance

f the dimensions are such that the effect of the conductor ıpe may be ignored, the two single-layer windings of a trans- mer can be considered to form the plates of a condenser ose capacitance, C_y, is given approximately by

$$C_y = \frac{10}{9}\pi p\epsilon\left(\frac{2a + 2d + t}{2t}\right) \times 10^{-12} \text{ farad} \quad . \quad . \quad (4)$$

is capacitance will be uniformly distributed. When the nension t becomes smaller than d the conductor shape can no ger be ignored, and for conductors of circular cross-section loser approximation can be obtained by multiplying eqn. (4) a factor F given by

$$F = \sqrt[4]{\frac{2t^2}{d^2}} \text{ arc tan } \sqrt[4]{\frac{d^2}{2t^2}} \quad . \quad . \quad . \quad . \quad (4a)$$

.3) The Product of Inter-winding Capacitance and Leakage Inductance

The maximum value of the in-phase component of the error to the last term in eqn. (2) can be obtained by combining ıs. (3), (4) and (4a), which gives

$$^2C_y(L_p + L_s) = \tfrac{1}{2}\omega^2\epsilon\left(\frac{\pi N_p}{3}\right)^2\left(\frac{t+d}{t}\right)(2a + 2d + t)^2F \times 10^{-20} \quad . \quad . \quad . \quad . \quad (5)$$

n practice, N_p, a, d and t may all vary over a wide range and s not feasible to examine eqn. (5) in a general manner. The ıension d, however, is usually determined by the current ing of the transformer. In the present application this rating 5 amp, for which a conductor diameter of 0·16 cm (16 s.w.g.) y be considered appropriate. The curves given in Fig. 2 n illustrate the manner in which $C_y(L_p + L_s)/\epsilon(\pi N_p/3)^2$ ies for different values of a and t when $d = 0{\cdot}16$ cm. In h case it will be noted that, as t is increased from a very small

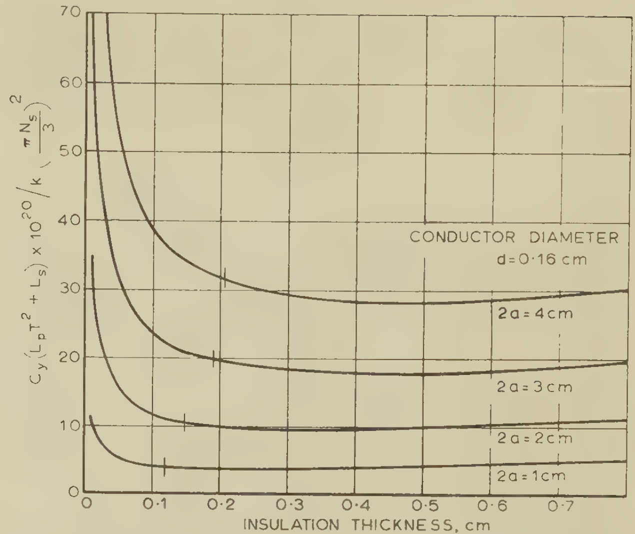


Fig. 2.—Change in the product of leakage inductance and inter-winding capacitance with variation of the insulation thickness.

quantity, the expression at first decreases rapidly and then becomes roughly constant. In the range where the error due to the last term of eqn. (2) is nearly independent of the value of t, the leakage inductance is increasing at about the same rate as the inter-winding capacitance is decreasing. Since the primary voltage drop at high frequencies is almost directly proportional to the leakage inductance and it is desirable that the former should be as low as possible, the smallest permissible value for t should be selected. It is suggested that this is given by values of t which correspond to 110% of the minimum of each of the curves in Fig. 2. The intercepts on these curves have been drawn accordingly, and it will be noticed that even over a 16 : 1 range in core area the optimum thickness of the insulation space varies only between 0·12 and 0·2 cm, a thickness not very different from the conductor diameter.

A series of curves for other values of a and d can readily be constructed, and using these in conjunction with eqn. (5) the maximum permissible number of turns, N_p, for any designed limit of error and frequency may be calculated. Fewer turns should not be used since this may adversely affect the errors at the lowest frequency of use or restrict the width of the working frequency band; reference to this will be made later. The size of core necessary to accommodate the required number of turns in a single layer can then be readily determined.

Several single-layer 5 amp-secondary transformers with differing numbers of turns and various core sizes have been constructed and their inter-winding capacitances and leakage inductances measured. In all cases the value of $C_y(L_p + L_s)$ calculated by means of eqn. (5) agreed with the measured product value to within 20%. The measured capacitances were, however, consistently lower and the leakage inductances consistently higher than their calculated values. This is due to the gaps which inevitably occur between adjacent turns around the outer circumference of the core and for which no allowance has been made. Better agreement can be obtained by multiplying the values of L_p and C_y calculated from eqns. (3) and (4) by empirical factors of 5/4 and 4/5 respectively.

(3.4) Inter-winding Capacitance and Circuit Resistance

The maximum value of the quadrature component of the error due to the last term in eqn. (2) is $j\omega C_y(R_p + R_s + 2R_b)/2$. Since the value of C_y in a given transformer is fixed by the considerations discussed in the previous Section the error is largely

controlled by the resistance of the windings. In multi-layer coils, skin and proximity effects cause the winding resistance to rise rapidly with the frequency, the effect being a function of the number of layers.[2] The effect is a minimum in a single-layer coil but even so may amount to a factor of two at 10 kc/s in a 5 amp conductor and must be allowed for.

(3.5) Self-Capacitance of the Windings

The two remaining high-frequency error terms, $Y_p(Z + Z_b)/T^2$ and $Y_s Z_b$, involve the self-capacitances of the windings. An isolated single-layer coil has a much smaller self-capacitance than the equivalent multi-layer coil since it has no layer-to-layer capacitances. The capacitance effective at the ends of a single-layer coil can be obtained approximately by using the well-known formula for round parallel wires to calculate the turn-to-turn capacitance and then dividing this result by the number of turns. It will be found that the self-capacitance is very small and usually it will be swamped by the capacitance of the leads used to connect the coil to its external terminals. In order to keep the lead capacitance small without introducing an unnecessary amount of inductance these leads must be kept as short as possible.

(3.5.1) The Primary Winding.

In a toroidal transformer the primary winding is usually the outermost and allowance must therefore be made for its capacitance to earth. Using the dimensions of Fig. 1 and assuming that the earth capacitance is uniformly distributed, its effective value at the terminals of the winding is $\pi\sqrt{(pa)}/18$ pF. The total self-capacitance of a single-layer primary winding (turn-to-turn plus leads plus earth capacitances) will always be many times smaller than the inter-winding capacitance, and consequently the error due to the term of eqn. (2) which involves Y_p will be small compared with that involving Y.

(3.5.2) The Secondary Winding.

To the self-capacitance of the secondary winding and its leads must be added the effective value of its capacitance to the conducting core. Knowing the thickness of the insulation the open-circuit capacitance between core and winding can be calculated using an equation similar in form to that of eqn. (4). Since the turns are wound in a single layer this capacitance is uniformly distributed and its effective value at the terminals is one-twelfth the open-circuit value. If the insulation space between core and winding is made comparable in thickness to that between the windings, then the error due to the term involving Y_s will also be small compared with that involving Y.

(3.6) Low-Frequency Errors

As the frequency is reduced, the remaining term of eqn. (2), $(Z_b + Z_s)/T^2 Z_c$, assumes an increasing importance, and at the lowest frequency of use the error of a well-designed transformer should be determined entirely by this term. Since it has been shown that the secondary leakage inductance of a single-layer toroidal transformer is negligibly small and non-inductive burdens are being considered, we may write $R_b + R_s$ for $Z_b + Z_s$. Using the dimensions of Fig. 1 it can then be shown that the frequency, f, and the phase error, γ, due to the reactive part of Z_c are related by

$$f = \frac{8 \times 10^7 p(R_b + R_s)}{\pi N_p^2 T^2 a^2 \mu_f \gamma} \quad . \quad . \quad . \quad . \quad (6)$$

where γ is in radians and μ_f is the permeability of the core at frequency f. Since all the quantities on the right-hand side of eqn. (6) except μ_f and γ have been fixed by the high-frequency requirements, the lowest working frequency for a given phase error is inversely proportional to the permeability of the c[ore]. Small errors at low audio frequencies therefore necessitate [the] use of those core materials which have very high values of in[itial] permeability.

The error due to the resistive component of Z_c results fr[om] the hysteresis and eddy-current losses in the core. Wit[h a] resistive secondary circuit the eddy-current losses will rem[ain] constant but the hysteresis losses will increase as the freque[ncy] is reduced; consequently, the error due to the combined lo[sses] will have its maximum value at the lowest frequency of [use]. The total loss in thin-strip high-permeability material is q[uite] small, and at the low levels of flux density required in a curr[ent] transformer no difficulty is experienced in reducing the ma[gni]tude error to a value well below that of the phase error gi[ven] by eqn. (6). In order to calculate the magnitude error i[t is] usual to consult the published curves which give total lo[sses as] functions of the flux density, frequency and strip thickness.

(3.7) Multi-Ratio Transformers

The conventional tapped primary winding having graded c[on]ductor sizes is unsatisfactory for wide-frequency-band operat[ion]. One of the many disadvantages lies in the fact that, at the hig[her] audio frequencies, the errors are certain to be different for ev[ery] ratio. The most efficient form of construction is to use all [the] winding turns for every ratio. If, therefore, the turns o[f a] complete single-layer primary winding are divided into secti[ons] which are reconnected by series-parallel arrangements, a mu[lti-] ratio transformer will be obtained which possesses al[l the] advantages of the single-layer form of construction. The im[pe]dances of the individual sections will be virtually identical, [and] consequently the errors of the various ratios will be sufficie[ntly] alike for a single calibration to be used.

The greatest restriction imposed by the series-parallel typ[e of] winding is on the number of ratios that conveniently may [be] provided. When, however, the operating conditions are sev[ere] the convenience of many ratios or the ease of changing them i[s of] secondary importance and the series-parallel method must [be] used.

(4) DESIGN PROCEDURE

The final design of most transformers which are require[d to] work over a wide band of frequencies is often a compro[mise] between the high- and low-frequency requirements, and trial-a[nd-] error methods may be necessary. In the present type of tra[ns]former the diameter of the conductors, d, and the resistanc[e of] the external burden, R_b, will be fixed by the current and po[wer] ratings, but there is a certain amount of latitude in the choice [of] the area of the magnetic core. Experience, however, has sho[wn] the need for at least 3 cm[2] but not more than 10 cm[2] of c[ore] area when precision accuracy is required at low audio frequenc[ies]. It is suggested, therefore, that an area of about 6 cm[2] should [be] taken for an initial design, i.e. the dimension $2a$ should be f[ixed] at 2·5 cm. The procedure then to be adopted for any step-do[wn] ratio transformer is first to determine the values of the remai[ning] parameters for the unity-ratio case in the following order:

(a) From the curves of Fig. 2 find the optimum thickness, [of] the interwinding space and decide upon a suitable dielectric mate[rial].
(b) Using eqn. (5) calculate the number of turns, N_p, for [the] specified error limit at the highest frequency of use.
(c) Calculate the mean magnetic radius, p, of a core which [will] accommodate $N_s(=N_p)$ turns in a single layer and deduce R_p and [R_s].
(d) Calculate from eqn. (6) the minimum value of the [initial] permeability, μ_f, required for the specified low-frequency error.
(e) Calculate L_p, C_y, C_p and C_s, determine the values of Z [at] both frequencies and then evaluate the remaining errors.

If the overall errors are now greater than specified, a sec[ond] trial with a suitably adjusted value for $2a$ will be necess[ary].

vill be realized that the limitations in the properties of existing e materials set a lower limit to the errors that can be obtained any specified frequency bandwidth. A worked example and 'able showing the relative values of the various errors are given Section 9.

When the final design details have been settled the required io or ratios can be obtained by subdividing the primary ıding into the appropriate number of sections and providing means for series-paralleling them.

(5) PRACTICAL ASPECTS OF DESIGN

a) *Core.*—Strip-wound clock-spring cores are now obtainable h an initial d.c. permeability of 45 000 ± 10%. Such high ues can only be maintained if the core is protected from chanical deformation and pressure during winding. This is st done by encasing the core loosely in a rigid plastic box of ll thickness about 0·1 cm. The use of hard core impreg-nts is prohibited, but soft greases may be used to limit any ovements.[3] The strip thickness should be in the range)05–0·01 cm in order to obtain an acceptable compromise tween the electrical and mechanical properties. Strip thicker in 0·01 cm will have unnecessarily high losses and a rapid fall permeability with increase in frequency, whilst cores with ip thinner than 0·005 cm and large enough to be of use for a rrent transformer are likely to be loosely coiled, have a poor ace factor and be difficult to handle.

b) *Windings.*—Synthetic-enamelled wire should be used for conductors. Silk or cotton coverings and all hygroscopic ulating materials are to be avoided. In order to prevent cessive capacitance between the end turns of any winding a all portion of the core (0·5 cm) should be left unwound, but gap in the primary winding must be in the same place ative to the core as the gap in the secondary winding. When minimum rated current of any primary section of a multi-io transformer is to exceed 5 amp, undue paralleling of the nductors can be avoided by using flat strip 0·16 cm thick tead of round wires. The width of such strip should not, wever, exceed about 1 cm, since it would then be difficult to nd closely over the inter-winding insulation and the resulting mary turns would be skewed with respect to the secondary ns.

c) *Inter-winding Space.*—The calculated thickness of the inter-ıding space is best realized by binding flexible strips and nular discs of polyethylene sheet into position by narrow thin lyethylene tape. This material has the advantage of low rmittivity and negligible power loss throughout the audio-quency band.

d) *Terminals and Terminations.*—The leads joining the main-rrent terminals to the ends of the primary windings should be ntinuations of the conductors wound on the core, arranged a system of identical go-and-return pairs disposed symmetri-ly with respect to the axis of the core. The length of the ds need not exceed 10 cm.

The spacing between the main-current terminals should be as all as possible in order to keep the series inductance and the istance of the series-parallel links to a minimum. Contact istance can be reduced by heavily silver-plating the terminals l links. For primary currents above about 100 amp the minals should be replaced by flat parallel busbars separated as little as 0·1 cm.

(6) CONSTRUCTION AND PERFORMANCE OF THE N.P.L. STANDARD TRANSFORMERS

Four standard multi-ratio transformers designed to have ors not exceeding 1 part in 10^4 over the frequency range 400 c/s–10 kc/s were constructed. All had 5 amp secondary windings, and between them rated primary currents from 5 to 400 amp were covered. Certain ratios were duplicated for the purpose of comparison tests.

(6.1) Constructional Details

(*a*) *Core.*—Clock-spring 15 cm o.d. × 10 cm i.d. of 0·01 cm × 2·5 cm 77/14H.P. nickel-iron alloy strip. Weight, 2 kg.

(*b*) *Secondary Winding.*—160 turns of 0·16 cm enamelled round copper wire wound in a uniformly distributed single layer. Resistances: 0·17 ohm (d.c.) and 0·4 ohm (10 kc/s).

(*c*) *Interwinding Insulation.*—Polyethylene sheet and film to a thickness of 0·18 cm.

(*d*) *Primary Winding.*—0·16 cm enamelled round wire or 0·64 cm × 0·16 cm enamelled strip wound in a uniformly distributed single layer and sectionalized for series-parallel connections. The main-current busbars of the 200 and 400 amp transformer were 6 cm × 0·6 cm in section separated by 0.1 cm. The inductance of these busbars was estimated to be 5 mμH.

The leakage inductance referred to the secondary and the capacitance between windings were measured and found to be about 50 μH and 200 pF respectively for each transformer.

(6.2) Performance

The ratios and phase angles of the transformers were measured by the bridge method which uses one resistance standard in series with the primary circuit and another in the secondary circuit. Unwanted electromagnetic and electrostatic couplings between the circuit components become increasingly important at high audio frequencies, and extreme care is necessary in order to eliminate errors in measurement due to their effects. The measurements are difficult and involve special detection equipment and new techniques. The details are outside the scope of this paper but have been described in full in a companion paper.[4] Table 2 gives the results obtained for all three ratios of one of the standard transformers, tested at its rated non-inductive burden of 5 VA. The phase angles are given in the more usual form of minutes of arc. The uncertainty in any value is estimated not to exceed 2 parts in 10^5 and 0·05′.

Table 2

Percentage of rated current	At 400 c/s	At 1 kc/s	At 3 kc/s	At 5 kc/s	At 10 kc/s
	True ratio/Turns ratio				
%					
120–10	1·000 04	1·000 04	1·000 04	1·000 03	1·000 02
	Phase angle				
%	min	min	min	min	min
120	+0·27	+0·12	+0·08	+0·04	+0·02
100	+0·30	+0·13	+0·08	+0·04	+0·02
60	+0·32	+0·14	+0·08	+0·04	+0·02
20	+0·34	+0·14	+0·08	+0·04	+0·02
10	+0·34	+0·14	+0·08	+0·04	+0·02

The maximum changes in error at 10 kc/s due to the connection of a low-impedance link between any primary and any secondary terminal were less than 3 parts in 10^5 and 0·03′.

The other transformers were found to have errors identical within one or two parts in 10^5 of those given in Table 2.

Tests were also made on a 5/5 ratio at frequencies outside the design range. At full-load current and 50 c/s the errors were +1 part in 10^4 and +1′, whilst at 30 kc/s they did not exceed ±1 part in 10^4 and ±0·1′ with any condition of primary to secondary linkage.

All the measured errors have been confirmed by calculation

using eqn. (2), and the agreement obtained was within the limits of the experimental uncertainties.

(7) ACKNOWLEDGMENTS

The work described has been carried out as part of the research programme of the National Physical Laboratory, and the paper is published by permission of the Director of the Laboratory.

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(9) APPENDIX

Design of a Precision Current Transformer

(*a*) *Specification.*—

Rated secondary current	5 amp.
Rated secondary burden	1 ohm, non-inductive.
Frequency range ..	400 c/s–30 kc/s.
Error limits	± 3 parts in 10^4 for magnitude and phase.

(*b*) *Procedure.*—Fix conductor diameter, core cross-section and permittivity of inter-winding insulation taking $d = 0{\cdot}16$ cm, $2a = 2{\cdot}3$ cm and $\epsilon = 2{\cdot}2$.

Assume the resistance of windings at 400 c/s to be the same as on direct current and at 30 kc/s to be three times as great.

From the appropriate curves and equations the following da are then obtained:

t (optimum insulation thickness)	$0{\cdot}17$ cm.
N_p	210
p (mean radius of core)	$6{\cdot}5$ cm.
R_p, R_s	$0{\cdot}23$ ohm at 400 c/s, $0{\cdot}7$ ohm at 30 kc/s.
μ_f (minimum permeability at 400 c/s)	29 000.
L_p	50 μH (corrected value).
C_y	280 pF (corrected value)
C_p	10 pF.
C_s	35 pF.

Taking the strip thickness of the core material to be $0{\cdot}01$ the following information is obtained from published data:

Total core loss at 400 c/s	$2{\cdot}75$ mW.
Total core loss at 30 kc/s	$2{\cdot}0$ mW.
Permeability at 30 kc/s	8 000.

The individual error terms of eqn. (2) calculated from t above values and the overall error of the transformer are g in Table 3.

Table 3

Error term	Value of error term, parts in 10^4	
	At 400 c/s	At 30 kc/s
$(Z_b + Z_s)/Z_c$	$0{\cdot}9 - j3{\cdot}0$	$0{\cdot}5 - j0{\cdot}2$
$Y_p(Z + Z_b)$	$0{\cdot}0$	$-0{\cdot}2$
$Y_s Z_b$	$0{\cdot}0$	$j0{\cdot}1$
Maximum value of last term, i.e. $Y(Z/2 + Z_b)$	$0{\cdot}0$	$-3{\cdot}0 + j0{\cdot}9$
Maximum overall error	$0{\cdot}9 - j3{\cdot}0$	$-2{\cdot}7 + j0{\cdot}8$

The overall errors of the transformer at the extreme frequencies of use are wi the specified limits.

[The discussion on the above paper will be found on page 337.]

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Sept. 1960

TECHNIQUES FOR THE CALIBRATION OF STANDARD CURRENT TRANSFORMERS UP TO 20 kc/s

By J. J. HILL, B.Sc., Associate Member.

paper was first received 11*th July*, 1959, *in revised form* 9*th January, and in final form* 24*th May*, 1960. *It was published in September,* 960, *and was read before the* MEASUREMENT AND CONTROL SECTION 10*th January, the* NORTH-EASTERN MEASUREMENT AND ELECTRONICS GROUP 16*th January, and the* NORTH-WESTERN MEASUREMENT GROUP 21*st February*, 1961.)

SUMMARY

ıen standard current transformers are tested at high audio encies the errors and uncertainties in the measurements are to be great owing to the effects of stray magnetic fields on the unshielded resistors which of necessity have to be used. The es of stray fields are examined and means are suggested for antially reducing them. It is shown that the 4-terminal resistance ards used in current-transformer testing bridges are very fre-cy-dependent and present techniques limit the accuracy of urement at 20 kc/s to 2 parts in 10^4 in ratio and $0\cdot 8'$ in phase. w techniques are described by which the errors and uncertainties e reduced to give an accuracy of 5 parts in 10^5 and $0\cdot 1'$ at 20 kc/s.

(1) INTRODUCTION

ıe theory and design of precision toroidally-wound current sformers has now been developed to the extent that their ɔrmance can be calculated to an accuracy better than 1 part $)^4$ in magnitude and phase for frequencies of use up to at 20 kc/s.[1] Transformers having overall calculated errors of v parts in 10^5 over the frequency range 400 c/s–10 kc/s have ıtly been constructed at the Laboratory, and it was essential the theoretical values should be verified experimentally.

t power frequencies the 'absolute' errors of a standard sformer are invariably determined by the null bridge method h uses two standard 4-terminal non-inductive resistors, one ıected in series with the primary and the other in series with econdary circuit. It was considered that this method would e satisfactory for audio frequencies, provided that tech-es were used which would eliminate the errors due to those ts which, though insignificantly small at power frequencies, ase proportionally with the frequency. The major difference een the current-transformer testing bridge and most other bridges is that all the main components are of very low dance and are large in size, and the working currents are high, frequently of the order of hundreds of amperes. The * and the high power dissipation of heavy-current resistance lards make efficient electromagnetic shielding a difficult expensive process and consequently it is rare for such ds to be fitted. If, therefore, the errors in measurement at audio frequencies due to unwanted e.m.f.'s induced in the ɔrs are not to exceed a few parts in 10^5, measures must ken to ensure that the stray magnetic fields produced by the currents in the circuit are reduced to an extremely low . A further source of error, and one not always appreciated, e degree of dependence upon frequency of the resistance ime-constant of 4-terminal standard resistors.

e following Sections deal with many aspects of precise urement of current transformers over a wide band of fre-cies and currents, and an account is given of a test procedure ı virtually eliminates all the errors and uncertainties of the lard method.

ıe 0·01-ohm 200 amp standard a.c. resistors used at the N.P.L. are about ng × 18 in high × 8 in wide.

paper is an official communication from the National Physical Laboratory.

(2) STRAY MAGNETIC FIELDS

The errors in measurement due to unwanted inductive couplings in the circuit are mainly caused by the stray magnetic fields of the primary-current supply leads and of the power transformer. The errors are proportional to frequency and consequently the precautions that are normally taken to avoid error at power frequencies may prove inadequate at audio frequencies. The problem of the reduction of these stray fields is closely linked with the problem of the transmission of large a.f. currents without excessive loss, and therefore the two aspects will be considered together.

(2.1) The Primary-Current Supply Leads

For a.f. currents greater than about 20 amp, twisted-pair circular-section conductors must not be used and should be replaced by a system having much smaller self-inductance and stray magnetic field. Concentric tubular conductors are not recommended, however, owing to the extreme difficulty of providing satisfactory end connections for the tubes, especially for the section of the circuit joining the standard resistor to the primary of the current transformer. Suitably low values of self-inductance and small external field can be obtained from go-and-return conductors arranged in the form of very thin flat strips with a small separation between them. This system is easy to construct and has a measure of flexibility not possessed by concentric tubular conductors, and the provision of suitable end connections presents no difficulty. A further advantage of thin strip conductors is that their large ratio of surface to cross-sectional area allows them to be operated at very high current densities without undue temperature rise. For example, two copper strips $\frac{1}{32}$ in × 2 in separated by 0·01 in of insulation are sufficient for a 200 amp go-and-return conductor system.

The self-inductance of this particular system is approximately $0\cdot 005\,\mu$H/ft, and consequently the frequency can be raised to 16 kc/s before the reactive voltage drop at rated current amounts to 0·1 volt/ft.

The small separation of the conductors which enables low values of self-inductance and stray field to be obtained leads, however, to large self-capacitance. Part of this capacitance will be shunted across the primary of the current transformer being calibrated and part across the standard resistor in series with it, and errors in measurement may therefore arise. The capacitance of the conductors referred to is approximately 1 000 pF/ft, and this would need to be shunted across 0·1 ohm in order to cause an error at 16 kc/s of 1 part in 10^5. Resistances and reactances as large as 0·1 ohm are only likely to occur in circuits where the maximum intended current is a few tens of amperes. In such cases much narrower conductor strip with smaller self-capacitance could be used. Experience with several parallel-strip systems has shown that errors due to the conductor capacitance can be made negligibly small.

(2.2) The Power Transformer

Even when parallel-strip conductors are used, the limited output available from most a.f. generators makes it necessary

for a heavy-current power transformer to be located within a few feet of the bridge components. In such circumstances it is essential for the transformer to be toroidally wound on a ring-shaped core. In order that the leakage reactances may be reduced as much as possible both the primary and secondary windings should be uniformly distributed, the terminal leads being arranged in close go-and-return pairs. A toroidal power transformer provided with four 3-volt 150 amp secondary windings has been constructed at the Laboratory and its external field is so small that it may be brought to within 3 ft of the circuit components before the bridge balance at 20 kc/s is disturbed by 2 parts in 10^5.

(3) STRAY CAPACITANCES AND EARTHING

Resistive leakage currents can be made negligibly small by using materials such as polyethylene for the insulation in all parts of the circuit. The transformer under test and the standard resistors can usually be arranged so that their mutual capacitances are as small as 10 pF whilst their earth capacitances are rarely as large as 50 pF. With stray capacitances of this order, errors in measurement are unlikely to be significant except in cases where the main current is less than about an ampere and where the circuit impedances exceed an ohm or so. In general it is not possible to use a Wagner earth on the bridge, on account of the low impedances in all the arms. It is, however, desirable to earth the bridge at one point, preferably where shown in Fig. 1. The mid-point of the primary of the power transformer may with advantage be earthed as well, but this is not always possible.

When tests are made over a wide band of audio frequencies the detector circuit usually consists of a high-impedance detector amplifier and an oscillograph, isolated from the main circuit by a screened balanced detector transformer. The precautions needed to minimize the effects of stray-capacitance currents in the detector circuit are well known, but further improvement can be obtained if the power supplies of the amplifier and oscillograph are also isolated from the 50 c/s mains supply by a low-admittance transformer. The magnetic shielding provided in the usual detector transformer may not be sufficient for the present application, and this transformer should, like others in the circuit, be toroidally wound.

(4) THE FREQUENCY CHARACTERISTICS OF LOW-VALUE 4-TERMINAL STANDARD RESISTORS

Four-terminal standard resistors intended for alternating currents are usually constructed using a bifilar round-wire or folded flat-strip resistance element. These elements can be designed so that their time-constants are as small as 0·1 microsec, but they are rarely rated to carry currents in excess of a few amperes. When larger currents are required, it is customary to connect the requisite number of elements in parallel by means of copper busbars to which the main-current and voltage terminals are attached, and the time-constant of such a resistor is often very different from that of the individual elements. The busbars can be regarded as a transmission line having inductance and resistance with a uniformly distributed resistive leak, and consequently the magnitude and phase of the voltage drop across the bars change as the voltage points are moved from one end to the other. It has been shown[2] that a resistor cannot be made non-inductive unless the inductance of the busbars is at least six times the inductive component of the impedance of the resistive elements. In a heavy-current resistor the busbars may be long and of considerable cross-section and the changes in the resistance and inductance of the bars as the frequency is varied will have a marked effect on the performance of the complete resistor. This phenomenon does not appear to have rec[...] much attention in the published literature. Formulae are [...]quently given for calculating the changes due to skin effe[ct in] individual bifilar wire or strip elements, but such changes [are] usually insignificant compared with those due to busbar eff[ects]. The a.c. resistance and inductance of the busbars is very de[pen]dent upon their geometry, and it is not usually possible to [cal]culate the performance of a complete resistor over a rang[e of] audio frequencies with sufficient accuracy. Therefore the r[esis]tance and time-constant of a standard resistor are always d[eter]mined by direct measurement. Table 1 gives the characteri[stics]

Table 1

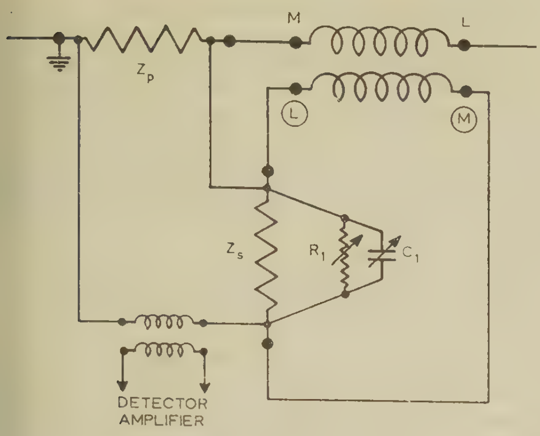
Frequency	Resistor A 0·4 ohm 5 amp 5 elements	Resistor B 0·1 ohm 20 amp 40 elements	Resistor C 0·01 ohm 200 amp 200 elements
	A.C./D.C. resistance ratio		
c/s			
50	1·0000	1·0000	1·0000
500	1·0000	1·0000	1·0000
1 000	1·0000	0·9999	0·9999
5 000	1·0001	0·9996	0·9991
10 000	1·0002	0·9994	0·9984
20 000	1·0006	0·9987	0·9979
	Time-constant		
	μs	μs	μs
50	+0·235	−0·006	+0·027
500	+0·235	−0·005	+0·038
1 000	+0·235	−0·004	+0·058
5 000	+0·235	+0·002	+0·108
10 000	+0·234	+0·007	+0·117
20 000	+0·233	+0·013	+0·124

of three resistors constructed with bifilar-wire elements, a[nd it] can be seen clearly that the frequency response is very depe[ndent] upon the number of elements connected in parallel.

For comparison, the changes over the whole frequency ra[nge] for a single-element resistor without busbars were less than 1 [part] in 10^4 and 0·001 microsec. The large changes shown in [the] Table demonstrate the necessity of determining the value[s of] standard resistors at every frequency of intended use. [Such] measurements are best carried out using Hartshorn's[3] modi[fica]tion of the Kelvin double bridge. Rayner[4] has recently in[tro]duced a further modification which eliminates one of the ca[uses] of uncertainty in the method. With a suitable choice of c[om]ponents the uncertainties in the measurements can now [be] reduced to 1 part in 10^4 in resistance and 0·001 microsec in t[ime-] constant for all frequencies up to 20 kc/s. Over a limited ra[nge] of resistance and provided the time-constants are small, [the] resistance ratio of two standards may be determined t[o an] accuracy of 5 parts in 10^5.

(5) EXPERIMENTAL DETAILS AND TEST PROCEDU[RE]

The complete bridge circuit is shown in Fig. 1 and the follo[wing] points of detail are to be noted. In order to ensure that [the] adjustable shunts R_1 and C_1 are always connected across [the] standard resistor Z_s the latter is constructed so that its resist[ance] is about 0·1% above the nominal value and its time-cons[tant] is about 0·1 microsec more positive than that of any v[alue] of Z_p. The difficulty of providing suitable shunts for Z_p w[hen] large primary currents and low impedances are involve[d is] thereby avoided. Z_s is usually between 0·1 ohm and 1 o[hm] and shunts consisting of a 4-decade capacitor sub-divide[d to] 100 pF and a 5-decade resistor subdivided to 0·1 ohm have [been] found to meet most practical requirements. If, when [the] detector deflection has been reduced to zero, the shunts o[f] have values of C_1 and R_1 and if R_p, R_s, l_p and l_s are the r



. 1.—Resistance-bridge method for the calibration of standard current transformers.
All voltage leads are arranged in closely twinned pairs.

es and residual inductances of Z_p and Z_s respectively, then ting terms in ω^4, the ratio of the transformer is given by

$$1 - \frac{R_s}{R_1} + \omega^2\frac{l_p}{R_p}\left(\frac{l_s}{R_s} - \frac{l_p}{R_p}\right) + \omega^2 C_1 R_s\left(2\frac{l_s}{R_s} - \frac{l_p}{R_p} - C_1 R_s\right)$$

$$- j\omega\left(1 - \frac{2R_s}{R_1}\right)\left(C_1 R_s + \frac{l_p}{R_p} - \frac{l_s}{R_s}\right)\Bigg] \quad . \quad (1)$$

re 'ratio' is defined as the quotient of primary current over ndary current and ω is 2π times the frequency. Eqn. (1) curate to 1 part in 10^5 at 20 kc/s provided the time-constants ne standard resistors do not exceed $0{\cdot}5$ microsec.

lthough every care may have been taken to make the various electric fields as small as possible it cannot be assumed their effects have, in fact, been reduced to negligible propor- s, and consequently a single bridge-balancing operation does provide sufficient data to determine the true value of the sformer ratio. The most likely error is that due to e.m.f.'s ced in the voltage circuits of the standard resistors. If only resistor were so affected the mean of two readings obtained reversed leads would theoretically be free from error. e, however, either or both resistors and the detector trans- er may be affected it is necessary to take pairs of readings, before and one after reversing the current and voltage leads $_p$, the current and voltage leads of Z_s and the primary con- ions to the detector transformer, and then taking the mean e of the eight results obtained.* If one side of the generator rthed it is desirable to repeat the procedure after reversing supply leads to the power transformer. It will not always ecessary to carry out this procedure in full, but only inspec- of actual results can indicate which reversals may safely be ensed with.

must be pointed out at this stage that when the above tests been completed the ratio of the transformer will have been sured for only one of the several conditions in which it be tested or used subsequently in practice. For example, e connections are made as shown in Fig. 1, then the M and erminals of the transformer are joined by a low-impedance and the ratio value would refer to that condition. The for the condition that L and Ⓜ are linked is determined eversing the leads to both the primary and secondary inals of the transformer and then repeating the test proce- dure. The ratio when L and Ⓛ or M and Ⓜ are linked or when there is no connection between the circuits cannot be measured directly without the use of special apparatus, and it is usual to calculate the values relating to these conditions from other, indirect, measurements.[5]

n general, this procedure will give the true result only if the effects of stray tance are small and unchanged by the reversal of connections. In the present e effects are negligibly small.

The accuracy with which the numerical values of the ratio and the phase angle of the transformer are finally obtained is entirely determined by the uncertainties in the values of the standard resistors and therefore at the worst may be no better than 2 parts in 10^4 and $0{\cdot}85'$. Whilst this accuracy is likely to be sufficient for most purposes it can be greatly improved upon if the standard transformer being tested has windings giving a nominal primary/secondary ratio of unity. In these circumstances it is possible to determine all the ratios of the transformer to an accuracy better than 5 parts in 10^5 and $0{\cdot}1'$ at all frequencies up to 20 kc/s without the need to know the precise values of the standard resistors. This method may be termed 'self-checking' or 'self-calibrating' and is described in Section 5.1.

(5.1) Self-Checking Method for Highest Accuracy

When the nominal ratio of the transformer under test is unity, the errors and uncertainties in the values of the standard resistors can be practically eliminated by repeating the test procedure after interchanging the resistors and then taking the mean of the results obtained. Inspection of eqn. (1), however, shows that in some circumstances not all the terms in ω^2 are completely eliminated and at very high audio frequencies a small residual error is left. If the errors of the transformer do not exceed 1 part in 10^3 in magnitude and $10'$ in angle and standard resistors having time-constants less than $0{\cdot}1$ microsec are used in the tests, the simplified form of eqn. (2) may be used to evaluate the true ratio without introducing errors greater than about one part in 10^5 at 20 kc/s.

$$\text{True ratio} = 1 - \frac{R}{2}\left(\frac{1}{R_1} + \frac{1}{R_1'}\right) - j\omega\frac{R}{2}(C_1 + C_1') \quad . \quad (2)$$

where R is the nominal value of the standard resistors and $1/R_1$, C_1 and $1/R_1'$, C_1' refer to the first and second sets of measurements respectively and are mean values obtained after carrying out the necessary reversals.

After the unity ratio of the transformer has been determined, it can be used to obtain a calibration of the circuit which will be required for tests on the next ratio. Since the principle of the following method is the same whether step-up or step-down ratios are subsequently involved, description will be confined to the latter case. Without altering the connections to the unity-ratio windings of the transformer, the existing primary standard resistor is replaced by one of value appropriate to the next ratio. In order now to be able to balance the bridge, means for obtaining the requisite fraction of the voltage drop across the secondary resistor has to be provided. There are many ways of doing this, but the author prefers to use a tapped voltage-dividing resistor connected across the voltage points of the secondary standard resistor. This divider, the constructional details of which are given in Section 8, has interchangeable sections of nominally equal value, giving a total resistance of 100 ohms. Special voltage leads are provided so that the total lead capacitance of the branch of the detector circuit that is connected to the pick-off section or sections is less than 2 pF. Consequently, for all practical purposes the errors of the voltage divider can be eliminated by repeating the bridge calibration for all the combinations of sections which give the desired voltage fraction and then taking the mean value. For example, whilst only two tests are required to obtain the true mid-point of a 2-section divider, ten tests with a 5-section divider are necessary for true fractions of 2/5 or 3/5 to be obtained.

When finally the mean bridge calibration has been determined, the primary of the transformer is altered to give the desired step-down ratio and the bridge is re-balanced using the full voltage across the divider. The results then obtained, after correcting for the bridge calibration, give the true value of the new ratio. The process can be extended indefinitely, using the determined value of each ratio in turn to provide a bridge calibration for the next.

It will be noted that the method described may involve a very large number of individual measurements, especially if the full procedure for the elimination of stray field errors is carried out at every stage. In general it will not be necessary to do this and frequently an experienced observer is able to reduce the number substantially below that theoretically necessary.

The principles and methods of test described have been used to determine the errors of a series of multi-ratio precision current transformers rated between 5/5 and 400/5, at frequencies from 400 c/s to 20 kc/s. The design of these transformers was such that their performance could be calculated to an accuracy of 2 or 3 parts in 10^5 and $0\cdot05'$. In all cases the agreement between the measured and calculated values was within the above limits.

(5.2) Comparative Method for Routine Testing

When calibrated standard transformers are available the routine testing of others can most easily be carried out by comparative methods. These have a great advantage over the 'absolute' or direct method in that only the difference in error between the standard and test transformers has to be measured and no heavy-current standard resistors are required. At least two types of self-contained current-transformer testing bridges working on this difference principle are available commercially and have been found satisfactory for their purpose up to the lower end of the a.f. band. For the frequency range covered in the present paper the simple bridge of Fig. 2 has

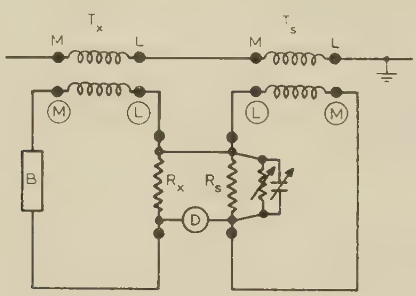


Fig. 2.—Comparative method for testing current transformers.
D = Detector.
B = Added burden.

been found to give excellent results. R_s and R_x are nominally equal non-inductive resistors designed to drop 1 volt at their rated current of 5 amp. Each is constructed from a single folded-strip element to which the main-current terminals are directly attached. The absence of busbars allows the strip to be folded back upon itself several times, making the complete unit small and very compact. The potential leads are very closely twinned with a minimum of insulation, and the potential terminals are arranged back-to-back with a separation of less than 1 mm. Tests have shown these resistors to be virtually unaffected by stray magnetic fields. It is not necessary for the absolute resistances and time-constants of the resistors to be known, the errors of the test transformer being determined in terms of the standard transformer by taking the mean of two sets of measurements in which R_s and R_x are interchanged.

Unlike the resistance-bridge method previously described, the comparative method enables the test transformer to be calibrated either without a direct link between the circuits or with a in any of the four possible positions. Since the capacitan between the primary circuit and both secondary circuits shared by the transformers, the calibration of the test tra former will be affected by the presence of the primary secondary capacitances of the standard, and vice versa. order that errors due to this cause should not arise, the de of the standard transformer should be such that its primary secondary capacitance is very small and that it is unaffecte the external circuit conditions.

Transformers having these qualities have been described companion paper[1] to which reference should be made.

(6) ACKNOWLEDGMENTS

The work described has been carried out as part of the resea programme of the National Physical Laboratory, and the pa is published by permission of the Director of the Laborat The author desires to acknowledge the assistance rendered his colleagues and in particular by Messrs. A. P. Miller G. H. Rayner.

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(8) APPENDIX

Voltage-Dividing Resistor

Each section of the divider consists of a 20-ohm resis bifilar-wound on a 1 in × 1 in × 0·01 in mica card, with 0·00 diameter synthetic enamelled constantan wire. The ti constants of individual sections were found to be less t 0·01 microsec, the differences between any two not exceed 0·001 microsec. The resistance cards are arranged in a ci of about 3 in diameter and are connected to a central circ terminal panel by twinned stiff copper wires to form a wheel structure. The terminal block is $\frac{1}{2}$ in diameter and is fitted $\frac{3}{8}$ in × $\frac{1}{16}$ in brass pins which serve to connect the section series and provide the voltage tapping points. Arrangem are made so that 4, 5 or 6 sections may be accommodated, giving voltage fractions covering all the ratios normally me current-transformer testing. All connections are made soldering. The input and output leads of the divider lie al the axis and on opposite sides of the plane containing resistance cards.

The voltage division was found to be unaffected by exte stray magnetic fields. The uncertainties in the absolute va of the voltage and time-constant ratios are eliminated from measurements by repeated tests in which all the indivi resistance cards in turn are used in the tapping section of divider. It should be noted, however, that the output vol must always be taken from a tapping point and one end.

DISCUSSION ON
'IMPROVEMENTS IN THE PRECISION MEASUREMENT OF CAPACITANCE'*
AND ON THE ABOVE THREE PAPERS
BEFORE THE MEASUREMENT AND CONTROL SECTION, 10TH JANUARY, 1961

rof. A. H. M. Arnold: For a long time the determination he ohm has been dependent on the standard of mutual ictance. Although improvements have been made in its struction, the accuracy of determination of the ohm has been ted to about one part in 100000, and perhaps not even that. technique of maintaining the ohm by means of tangible dards is very good, and the uncertainty arising from drift a period of 20 years may be less than one part in 100000, hat a new absolute determination after this interval of time ely results in confirmation, to a lower accuracy, of a value ady known. There is now a possibility of determining the 1, using a standard of capacitance, to an accuracy within ut one part in 10^6. Should this be achieved, the accuracy stability of tangible standards will need to be improved espondingly. It is possible that the alloy, Evanohm, rred to in the paper by Messrs. Harkness and Wilkins, may ve useful for this purpose, provided that its stability can be blished beyond suspicion.

he transformer constructed by Messrs. Hill and Miller has markable performance over a wide range of frequency, but might well have been made of the very fine digital computer he National Physical Laboratory in order to optimize the gn. The objective should be to design a transformer in ch the minimum error occurs in the middle of the working uency range, and this has not been achieved by the authors. reover, the poor use made of the available winding space has lted in a transformer which is larger than necessary, and this t have an adverse effect on both the magnitude of the gnetizing current and the magnitude of capacitance effects. s quite possible that a detailed calculation would show that yer windings with a correspondingly reduced length of gnetic path would result in a better overall performance than actually obtained. An alternative approach is either to use ductors of non-circular cross-section or to use parallel-nected windings for the secondary.

also should the thickness of insulation between the secondary winding and the core.

It is further quite unjustifiable to assume that the leakage reactance of the inner winding is necessarily near to zero. When a transformer has a multi-layer inner winding its leakage reactance may be near zero or even slightly negative, but with a single-layer winding a significant positive leakage reactance is possible.

Mr. Hill, in his paper, has a curious method of defining the external field of the power transformer. Could it not be defined in a more orthodox manner?

These minor defects in the four papers from the National Physical Laboratory should not be allowed to obscure the fact that a very real contribution has been made to the art of accurate measurement in the audio-frequency range.

Mr. A. C. Lynch: The valve amplifier, used as described, must meet stringent conditions. Would the design problem be eased if it had been used instead as part of a null indicator for comparing the unknown voltage with one derived by division on potentiometers from a voltage large enough to be measured with the standard electrometer?

There is mention of the use of a transformer-coupled bridge, of ratio 10 : 1, for making a chain of comparisons between ordinary standards and the very small capacitances obtained with the Thompson-Lampard type of standard. Since these small 3-terminal capacitances are perfectly additive, would it not be more reliable to build up large capacitances from small ones by substitution methods alone?

Mr. A. Cooper: In the paper by Messrs. Harkness and Wilkins the amplifier feedback to the input terminals is taken from the cathode resistor to the grid resistor of V_1. Therefore, as the authors point out, the drive impedance affects the performance. It may cause self-oscillation if sufficiently inductive, or, even more serious, it may itself be affected in such a manner as to alter considerably the voltage across it.

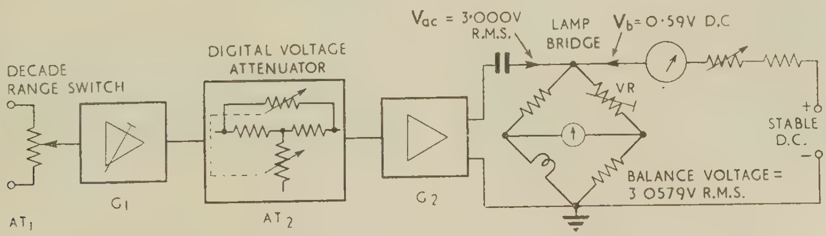

Fig. A.

am quite unable to accept the authors' proposition that the section is fixed. Both its magnitude and shape should be sidered as variables in the optimizing calculations, and so

I should like to give details of a development of the lamp bridge which is now incorporated in a commercial precision voltmeter covering the range 1 mV–300 volts and 5 c/s–100 kc/s.

In Fig. A the decade range switch AT_1 limits the voltage range to be handled by the amplifier G_1 to one decade. The novel

AYNER, G. H., and FORD, L. H.: Paper No. 3159 M, March, 1960 (see 107 B, 5).

digital voltage attenuator AT_2 has four decade dials covering a total of one decade, giving a minimum setting accuracy of 0·1%. This device feeds an amplifier G_2 which energizes the lamp bridge, which is a true r.m.s. indicator. The bridge is also fed with direct current and arranged to balance when the r.m.s. (heating) value of the a.c. signal plus d.c. level reaches a predetermined level. The direct current enables a highly sensitive spot galvanometer to be used for balance, independent of frequency.

3 volts a.c. and 0·59 volt d.c. have been chosen, so that the direct current has an effect of only about 2% on the balance, whilst the total of 3·0579 volts r.m.s. enables standardization of the bridge to be made indirectly against three Weston industrial standard cells.

The bridge is then used to standardize a stable a.c. source to 3 volts, which is divided down to 100·00 mV, and, in turn, applied to the system input and used to standardize it by adj[usting] the gain of the amplifier G_1.

A discrimination of 0·01% at all voltages is readily obtaina[ble] with errors not exceeding 0·05% over a wide frequency ra[nge].

Dr. T. G. Hammerton: In Section 4 of the paper by Mes[srs.] Harkness and Wilkins we have the unusual statement:

> As the instrinsic value of both the instrument and the work [in] which it will be employed is high, complete and fairly frequent v[alve] replacement will be justified. Thus with replacement after [no] more than 1 000 hours' use very little deterioration of properti[es is] expected.

This is the reverse of what I should have expected. It is [not] unusual for new communication-type valves to be 'soaked' [for] at least 1000 hours, during which time their characteri[stics] change relatively enormously and rapidly. We then find 'a[ged'] valves to be reasonably reliable, long lived and constan[t in] characteristics.

THE AUTHORS' REPLIES TO THE ABOVE DISCUSSION

Messrs. G. H. Rayner and **L. H. Ford** (*in reply*): In reply to Mr. Lynch, small 3-terminal capacitors, if of sufficient stability, provide an excellent means of checking a transformer ratio. However, when it is desired to compare capacitors differing in value by several decades it is more accurate and convenient to use a calibrated 10 : 1 transformer than to build-up over a wide range using only capacitors; the transformer will be more stable than the capacitors, and fewer capacitors and much less time are required.

Messrs. S. Harkness and **F. J. Wilkins** (*in reply*): Our paper is concerned only with giving an account of a precision amplifier designed for the specific purpose of increasing the sensitivity of the a.c./d.c. transfer instruments used at the N.P.L. No doubt there are other ways in which small alternating voltages may be measured. Mr. Lynch's proposal, whilst allowing the use of an amplifier of inferior accuracy as far as gain is concerned, raises a new set of problems, e.g. the provision of a reference source of high-voltage stability and having the same phase and frequency as the voltage being measured.

With regard to Mr. Cooper's point about the amplifier being likely to oscillate when the source is sufficiently inductive, this is avoided by the inclusion of sufficient series resistance in the input circuit. By proper design this loss can be made small enough, in all practical cases, to have negligible effect on the circuit performance.

It is agreed, as Dr. Hammerton points out, that there will be a fairly rapid change in valve performance, at least during the first few hundred hours, and undoubtedly some valve ageing occurred during the experimental stages of the work. The accepted life of the domestic-type valves used is only 3 000–5 000 hours, and replacement at about 1 000 hours is suggested in order to reduce the probability of failure of the equipment during use.

Messrs. J. J. Hill and **A. P. Miller** (*in reply*): We agree with Prof. Arnold that, ideally, the minimum error of a transformer should occur in the middle of its working frequency range. [This] requirement is, in fact, achieved in our transformers if the ra[nge] is taken as 400 c/s to 30 kc/s: it was always intended that [the] original range of 400 c/s to 10 kc/s should be extended if possi[ble].

Whilst smaller l.f. errors would result from the use of 2-la[yer] windings with a correspondingly reduced magnetic path len[gth], it is probable that the h.f. errors would be increased. Since [the] present demand is for greater accuracy and certainty of meas[ure]ment at high audio frequencies, improvement at the l.f. en[d is] of minor significance.

Reference is made to the possible use of non-circular c[on]ductors, and it is suggested that edgewise-applied interlea[ved] wide strips would make better use of the available win[ding] space. We cannot recommend this since its use leads to extre[mely] large capacitances. Moreover, the fact that the strips hav[e to] be individually shaped and soldered together on the core m[akes] the process laborious and costly. The National Bureau [of] Standards have, however, recently adopted this form of c[on]struction in a 5/5 current transformer designed for use u[p to] 10 kc/s. The resulting interwinding capacitance is stated t[o be] 40 000 pF, and the errors, which already have become nega[tive] at about 2 kc/s, are at 10 kc/s several times as great as in [the] N.P.L. transformers.

We had not intended to imply that the core section is invaria[nt], and the curves in Fig. 2 of the paper illustrate the effect o[f] variation. The wall thickness of the necessary protective [box] for the core is usually sufficient to make the effect of [the] capacitance between core and the secondary winding i[nsig]nificantly small.

Finally, with reference to the secondary leakage reactanc[e of] single-layer windings, we can only state that detailed exp[eri]mental work on several transformers showed that the secon[dary] winding contributed 5% or less to the total leakage react[ance] and that there was some evidence of the values being nega[tive] in sign.

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CIRCUIT FOR REDUCING THE EXCITING CURRENT OF INDUCTIVE DEVICES

By D. L. H. GIBBINGS, B.E., B.Sc., Ph.D.

(*The paper was first received* 10*th May, and in revised form* 1*st December*, 1960.)

SUMMARY

negative feedback device is described in which the greater part of xciting current necessary to energize a magnetic core is supplied n amplifier, so that windings on the core present a much higher edance than with the core material alone. An improvement of than 100 times has been achieved at 50 c/s. To the external it, the device appears as a composite core which may be wound in desired manner.

n investigation is made of the effect of using such a core on the ge and current ratios of a 2-winding transformer, and a proposal ade for using a shielding technique to achieve high precision.

LIST OF PRINCIPAL SYMBOLS

N = Number of turns.
A = Area of cross-section.
μ = Complex permeability.
G = Amplifier voltage gain.
Z_{in} = Amplifier input impedance.
Z_{out} = Amplifier output impedance.
, z_{22} = Self-impedances of windings 1 and 2.
, z_{21} = Mutual impedances of windings 1 and 2.
z_l = Load impedance.
z_s = Source impedance.

(1) INTRODUCTION

eal transformers could be made to approach more closely the l transformer were it not that real magnetic cores require ing currents to establish flux in them. By the use of the -permeability core materials now available, exciting currents be reduced to an extent which is adequate for ordinary irements, but for more critical applications a further reduc- is still valuable, and numerous methods have been described e literature[1–3] for bringing it about.

ntil this paper was being prepared, however, the author unaware of any proposal, other than his own, which was not tially a compensation scheme, the reduction in exciting ent depending on an impedance balance of some sort. It the difficulty of making a compensation scheme work ciently well over a band of frequencies and voltage loads h initiated the author's search for a negative feedback me as an alternative.

ne exception to the general use of compensation noted above proposal made in a Bulletin of the National Research ncil of Canada[4] which also employs negative feedback. It , however, suffer from some unnecessary limitations imposed he use of feedback round a single core, from which the ent scheme is free, and a description of the latter will be n.

rst there is a description in physical terms of the mode of ation of the device in magnifying the impedance of a simple ctive winding. Circuit details and results from an experi- cal model are given. Then follows a discussion of the errors 2-winding transformer and their reduction by means of such a circuit. It is shown that the combination of the present technique and a shielding technique recently described leads to the possibility of a transformer of extreme precision.

tten contributions on papers published without being read at meetings are l for consideration with a view to publication.
Gibbings is in the Division of Electrotechnology, National Standards Labora- Commonwealth Scientific and Industrial Research Organization, Australia.

(2) DESCRIPTION

When an alternating voltage is applied to a winding on a magnetic core, the exciting current that flows provides just sufficient magnetomotive force to establish the flux, and hence the induced voltage necessary to balance the applied voltage minus the drop in the winding resistance. If one can arrange to generate a proportion of the flux by some means which does not load the source, the exciting current will be correspondingly reduced and the apparent impedance of the winding raised.

In the present device (Fig. 1) the magnetic core is divided into

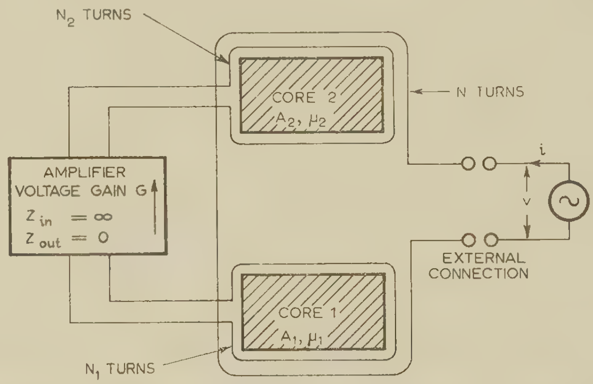

Fig. 1.—Circuit for reducing the exciting current of a winding.

two parts, which are separately wound and connected by an amplifier. The winding whose impedance is to be magnified links both cores.

When a voltage is applied to this winding, the sum of the fluxes in cores 1 and 2 adjusts itself to the applied voltage as before, but the amplifier ensures that the flux in core 2 is GN_1/N_2 times that in core 1, and supplies the exciting magnetomotive force to establish it, leaving only the flux in core 1, $1/(1 + GN_1/N_2)$ of the total, to be established by the current drawn from the external source. The winding linking the two cores thus appears to the external source to have a higher impedance than it would if the amplifier were not there.

To obtain a more quantitative picture of the operation, suppose that one applies a voltage v to the N-turn winding. The flux in the two cores must equal K_1v/N, where K_1 is a constant depending on frequency.

If, for simplicity, one assumes equal magnetic path lengths in the two cores, the exciting current, i, in the absence of the amplifier, establishes a flux proportional to $Ni\mu_1A_1$ in core 1 and $Ni\mu_2A_2$ in core 2, such that the total flux is

$$K_2Ni(\mu_1A_1 + \mu_2A_2) = K_1\frac{v}{N}$$

where K_2 is a constant depending on the magnetic path length.

Thus the N-turn winding has an impedance

$$z_N = \frac{K_2}{K_1}(\mu_1 A_1 + \mu_2 A_2)N^2$$

z_N is simply the sum of the separate impedances z_{N1} and z_{N2} of N turns on each core, coupling between cores being assumed negligible.

With an amplifier present,

$$K_1\frac{v}{N} = K_2\left(Ni\mu_1 A_1 + G\frac{N_1}{N_2}Ni\mu_1 A_1\right)$$

so that the N-turn winding now has an impedance

$$Z_N = \frac{K_2}{K_1}\mu_1 A_1 N^2\left(1 + G\frac{N_1}{N_2}\right) \quad . \quad . \quad . \quad (1)$$

resulting in an impedance magnification of

$$\frac{Z_N}{z_N} = \frac{\mu_1 A_1}{\mu_1 A_1 + \mu_2 A_2}\left(1 + G\frac{N_1}{N_2}\right) \quad . \quad . \quad . \quad (2)$$

It will be seen from eqn. (1) that the magnified impedance is $1 + GN_1/N_2$ times the impedance, Z_{N1}, of the same winding on core 1 alone.

Minor modifications to the analysis are needed if the amplifier is not an ideal voltage amplifier with infinite input impedance and zero output impedance. If the input impedance, Z_{in}, is finite, Z_N becomes $(1 + GN_1/N_2)$ times the resultant of z_{N1} shunted by $(N/N_1)^2 Z_{in}$. The output impedance of the amplifier has only a second-order effect; even if it were infinite, Z_N would be increased only by z_{N2}, the impedance of the N-turn winding on core 2.

It will be noticed that the amplifier effectively separates the requirements that a core material should have high permeability to give high impedance and a high saturation flux density to give a high maximum voltage per turn. With high gain, the impedance is determined almost entirely by core 1 and the voltage rating by core 2, and it may be advantageous to make the cores of different materials or to adjust their relative cross-sections to meet a particular specification.

The technical difficulties of the device, regarded as an exercise in feedback-amplifier design, reside in the fact that the external source impedance, which is not under the designer's control, appears in the feedback loop between two transformers (Fig. 2).

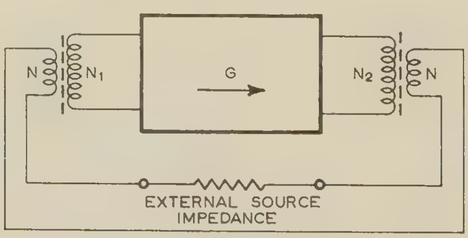

Fig. 2.—Exciting-current reduction circuit drawn as a feedback amplifier.

A capacitive source impedance would by itself introduce an asymptotic phase shift of π radians at low frequencies. The circuit of Fig. 3 has been designed to be stable with inductive or resistive source impedances of any magnitude.

To control the behaviour at high frequencies the gain is reduced by the internal feedback capacitor C_1, and another capacitor, C_2, is shunted across the terminals of the device. Where this generality of source impedance is not required, the design may be modified to extend the operating frequency band upwards.

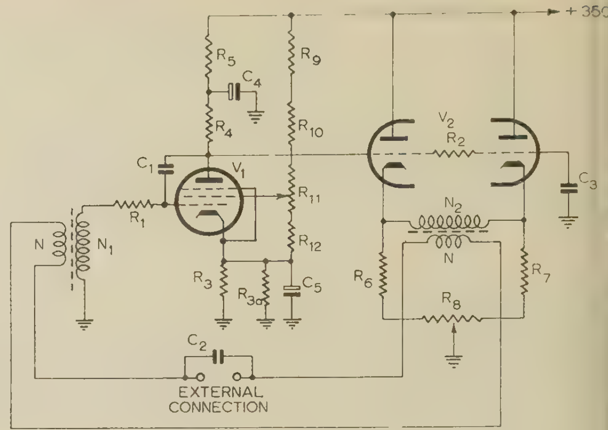

Fig. 3.—Experimental circuit.

C_1	100 pF	R_1	10 kΩ
C_2	4700 pF	R_2	2·2 MΩ
C_3	0·5 μF	R_3	330 Ω
C_4	8 μF	R_{3a}	1 kΩ
C_5	500 μF	R_4	470 kΩ
V_1	EF86	R_5	47 kΩ
V_2	12AT7	R_6, R_7	10 kΩ
N	22 turns	R_8	1 kΩ potentiometer
N_1, N_2	200 turns	R_9	56 kΩ
		R_{10}, R_{12}	6·8 kΩ
		R_{11}	10 kΩ potentiometer

The 22-turn windings shown separately actually form a single winding link[...] cores. The cores are $2 \times 1\frac{1}{4} \times \frac{3}{8}$ in Supermumetal toroids.

It is to be noted that C_2 is the only component which [...] to be changed when the number of turns in a winding is cha[...] Otherwise the device appears externally as a composite [...] drawing a greatly reduced exciting current.

(3) EXPERIMENTAL RESULT

Table 1 gives the results of measurements made on the 22[...] winding of Fig. 3 at five frequencies. For comparison, fi[...] are given for the impedance of 22 turns on two similar [...] without the amplifier.

Table 1

Frequency	Impedance with amplifier	Impedance without amplifier	Magnification
rad/s	Ω	Ω	
$3{\cdot}3 \times 10^2$	1800	14	130
$1{\cdot}0 \times 10^3$	6500	42	155
$3{\cdot}3 \times 10^3$	12000	140	86
$1{\cdot}0 \times 10^4$	5200	420	12
$3{\cdot}3 \times 10^4$	1600	1400	1

It will be seen that an impedance magnification of app[...]mately 100 is maintained up to 3300 rad/s (525 c/s), but [...] thereafter the combined effects of the stray capacitance [...] transformer windings, of C_2 and of the feedback capacit[...] cause the impedance to fall, so that after 33000 rad/s (525[...] the amplifier is actually disadvantageous. This result serv[...] emphasize one limitation of the scheme—that the ind[...] feedback does nothing to reduce the effects of capaci[...] currents in a winding. This does not mean that such a sc[...] is unworkable at high frequencies, but it does limit the frequ[...] band over which a single transformer will work.

It has been suggested to the author that the capaci[...] loading could be balanced out by a further winding linkin[...] two cores in the opposite sense with a capacitor in s[...] Indeed, the negative of any impedance could, in principl[...] produced in a similar way, the amplifier serving mere[...]

ve the unwanted shunting effects of the transformer. her work may be undertaken in this direction.

(4) TWO-WINDING TRANSFORMER

e previous Section has been concerned with magnification e impedance of a single winding on a composite core, and as for an application such as this—to improve the ratio of mpedance to winding resistance in an auto-transformer ge divider—that the idea was first developed. But in a ge divider one is also interested in ratio accuracy, and this ct will now be treated, using the voltage and current ratios 2-winding transformer as a more general case. The circuit tions for two windings on a core, with self-impedances z_{11}

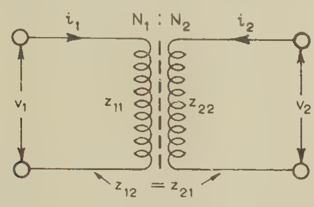

Fig. 4.—Two-winding transformer: nomenclature.

z_{22} and mutual impedances z_{12} and z_{21} (Fig. 4), are well wn to be

$$v_1 = z_{11}i_1 + z_{12}i_2 \quad . \quad . \quad . \quad . \quad . \quad (3)$$

$$v_2 = z_{21}i_1 + z_{22}i_2 \quad . \quad . \quad . \quad . \quad . \quad (4)$$

n equivalent form for these equations[5] is

$$v_1 = \left(z_{11} - \frac{N_1}{N_2}z_{12}\right)i_1 + \frac{z_{12}}{N_2}(N_1i_1 + N_2i_2) \quad . \quad (5)$$

$$v_2 = \frac{z_{21}}{N_1}(N_1i_1 + N_2i_2) + \left(z_{22} - \frac{N_2}{N_1}z_{21}\right)i_2 \quad . \quad (6)$$

in the case of a passive transformer for which $z_{12} = z_{21}$, are readily seen to be the equations for the equivalent circuit ig. 5, where the real transformer is replaced by an ideal one

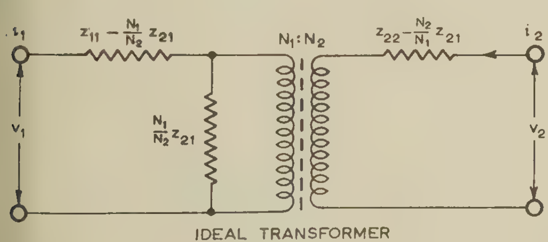

Fig. 5.—Two-winding transformer: equivalent circuit.

he same turns ratio. The series or leakage impedances esent the winding resistances plus the impedance due to which links one winding only and thus contributes to the but not the mutual impedances. Because this leakage must necessarily follow a path whose reluctance is determined ely by air, the leakage impedances are nearly independent of core material. The shunt impedance, on the other hand, is rted to account for the exciting current of the transformer depends directly on the core material.

is clear from the diagram that even with the secondary n-circuit ($i_2 = 0$) and a zero-impedance source ($v_s = v_1$), the voltage drop in the primary leakage impedance due to the exciting current causes the voltage ratio to become

$$\frac{v_s}{v_2} = \frac{z_{11}}{z_{21}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

rather than N_1/N_2.

Similarly, even with a perfect short-circuit load impedance ($v_2 = 0$), the voltage drop in the secondary leakage impedance causes some of the primary current to flow down the shunt path to excite the core, and the current ratio becomes

$$\frac{i_1}{i_2} = -\frac{z_{22}}{z_{21}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

If the source impedance is z_s rather than zero, but still with an open-circuit secondary ($i_2 = 0$), the voltage ratio is

$$\frac{v_s}{v_2} = \frac{z_{11} + z_s}{z_{21}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

and the analogous current ratio (for a load impedance z_l) is

$$\frac{i_1}{i_2} = -\frac{z_{22} + z_l}{z_{21}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

The loading of the secondary circuit with an impedance z_l, with zero source impedance, causes an additional voltage drop in both series impedances, and leads to a voltage ratio of

$$\frac{v_1}{v_2} = \frac{z_{11}}{z_{21}}\left(1 + \frac{z_{11}z_{22} - z_{12}z_{21}}{z_{11}z_l}\right) \quad . \quad . \quad . \quad (11)$$

The analogous current ratio, which corresponds to the case of a current source of finite shunt impedance z_s and zero-impedance secondary load, is

$$\frac{i_1}{i_2} = -\frac{z_{22}}{z_{21}}\left(1 + \frac{z_{11}z_{22} - z_{12}z_{21}}{z_{22}z_s}\right) \quad . \quad . \quad . \quad (12)$$

As before, $z_{11}z_{22} - z_{12}z_{21}$ is not zero because of imperfect coupling and winding resistance.

To make $(z_{11} + z_s)/z_{21}$ a good approximation to N_1/N_2 requires that

$$\frac{N_1}{N_2}z_{21} \gg z_{11} + z_s - \frac{N_1}{N_2}z_{21}$$

Similarly, for $(z_{22} + z_l)/z_{21}$ to approach N_2/N_1 requires that

$$\frac{N_2}{N_1}z_{21} \gg z_{22} + z_l - \frac{N_2}{N_1}z_{21}$$

Thus the direct way to reduce ratio errors is seen to be to increase the core permeability, since this has the effect of increasing z_{21} relative to $z_{11} - (N_1/N_2)z_{21}$ and $z_{22} - (N_2/N_1)z_{21}$. Whether the relative increase is used to increase z_{21} absolutely, leaving the series impedances unchanged, or to obtain the same value of z_{21} with fewer turns of heavier wire and hence smaller series impedances, is an engineering compromise depending on the principal source of the errors implied in eqns. (9), (10), (11) or (12).

By the previous physical reasoning, the improvement in ratio that occurs with increased core permeability can be explained in terms of a decrease in exciting current, but it cannot be immediately assumed that a reduction in exciting current achieved by artificial means, such as a compensation or a feedback scheme, will have precisely the same effect, and a closer analysis of a 2-winding transformer wound on a core whose exciting current

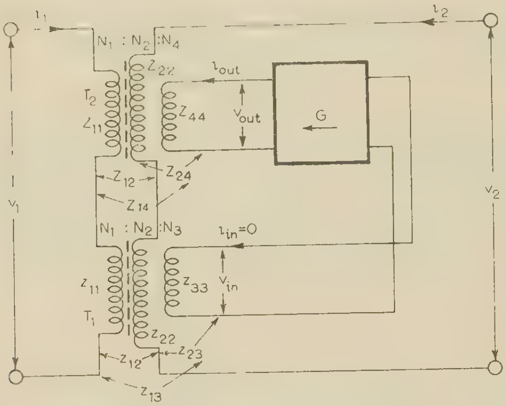

Fig. 6.—Two-winding transformer with exciting-current reduction.

is reduced by the present feedback scheme will now be made. The circuit equations applying to Fig. 6 are

$$v_1 = (Z_{11} + z_{11})i_1 + (Z_{12} + z_{12})i_2 + Z_{14}i_{out}$$
$$v_2 = (Z_{21} + z_{21})i_1 + (Z_{22} + z_{22})i_2 + Z_{24}i_{out}$$
$$v_{in} = z_{31}i_1 + z_{32}i_2$$
$$v_{out} = Gv_{in} = Z_{41}i_1 + Z_{42}i_2 + Z_{44}i_{out}$$

Eliminating v_{in}, v_{out} and i_{out} leads to two equations analogous to eqns. (3) and (4):

$$v_1 = \left[\left(Z_{11} - \frac{Z_{14}Z_{41}}{Z_{44}}\right) + z_{11} + G\frac{Z_{14}z_{31}}{Z_{44}}\right]i_1 + \left[\left(Z_{12} - \frac{Z_{42}Z_{14}}{Z_{44}}\right) + z_{12} + G\frac{z_{32}Z_{14}}{Z_{44}}\right]i_2 \quad (13)$$

$$v_2 = \left[\left(Z_{21} - \frac{Z_{24}Z_{41}}{Z_{44}}\right) + z_{21} + G\frac{Z_{24}z_{31}}{Z_{44}}\right]i_1 + \left[\left(Z_{22} - \frac{Z_{24}Z_{42}}{Z_{44}}\right) + z_{22} + G\frac{Z_{24}z_{32}}{Z_{44}}\right]i_2 \quad (14)$$

It will be noticed first of all that, because of the presence of active circuit-elements, the two mutual-impedance coefficients of eqns. (13) and (14) are no longer equal, although if the two subtransformers are of good quality, the difference will not be large, since

$$G\frac{z_{32}Z_{14}}{Z_{44}} \simeq G\frac{Z_{24}z_{31}}{Z_{44}}$$

i.e.

$$\frac{z_{32}}{z_{31}} \simeq \frac{Z_{42}}{Z_{41}}$$

Because of the inequality of the mutual impedances, an equivalent circuit such as Fig. 5 cannot be drawn. The terms $Z_{11} - Z_{14}Z_{41}/Z_{44}$, etc., will also be small if transformer 2 is of good quality; these terms appear in the equation because of the assumption of zero output impedance for the amplifier.

If $G = 0$, the voltage and current ratios, for ideal source and load conditions, approximate to z_{11}/z_{21} and $-z_{22}/z_{21}$ as before, but as G is made progressively larger, comparison of eqns. (13) and (14) with eqns. (9) and (10) shows that the ratios tend not to the turns ratio, as with a true increase in permeability, but to Z_{13}/Z_{23} and $-z_{23}/z_{13}$, respectively. That is to say, the effect of decreasing the exciting current by the feedback technique is to cause the voltage and current ratios of the composite transformer to approach the ratios of the mutual impedances of the transformer windings to the third winding on the output and input core, respectively. A result comparable with this see[ms] to be a necessary consequence of any scheme for reducing [the] exciting current by energizing the core by current in a th[ird] winding. One would expect the ratio of two mutual impedan[ces] to be closer than the ratio of a self- and a mutual impedance [to] the turns ratio, even without special adjustment, though [it] should be observed that it is the true permeability of the c[ore] material and not an artificially enhanced value which is import[ant] in fixing the mutual impedances. It is, of course, possible [to] adjust the ratio of the mutual impedances to be equal to [the] turns ratio.

The exact calculation of the effect of a secondary load on [the] voltage ratio of the composite transformer after the manner [of] eqn. (11) is made clumsy by the large number of terms in [the] coefficients of eqns. (13) and (14), but it is found to approxim[ate] to the effect of loading a transformer which has a very h[igh] shunt impedance and leakage impedances equal to those o[f the] two subtransformers in series, as one might have exp[ected] intuitively.

The most important result so far in the investigation o[f the] ratio of a 2-winding transformer wound on the composite c[ore] is that, as the gain in the feedback loop is increased, this ra[tio] tends to that of the mutual impedances between each princi[pal] winding and a subsidiary winding. The result is of importa[nce] because Thompson[6] has described a shielding technique by wh[ich] the ratio of the mutual impedances between two windings a[nd a] third can be made equal to the ratio of the turns of the two wi[nd]ings to a very high order of precision (greater than 1 in 1[0...]). The present result shows that, with sufficient gain, si[milar] precision can be achieved for the ratio of a 2-winding transfo[rmer].

In Thompson's transformer, the shielding technique [is] applied to make the ratio of the voltages induced in two second[ary] windings by a current in a primary one equal to their turns ra[tio] and may thus be taken over directly for T_2 (Fig. 6), the subtra[ns]former driven from the amplifier output. The driving windi[ng], N_4, goes inside the shields and the secondaries are formed by [the] turns N_1 and N_2 of the principal windings. With T_2 shiel[ded] in this way, the voltage ratio v_1/v_2 is made precise.

Because of the reciprocity of mutual impedance, it is a[lso] possible to use the same shielding technique to arrange t[hat] equal magnetomotive forces in two windings outside the shie[lds] induce equal voltages in a third winding inside. This is exa[ctly] what is desired for T_1 (Fig. 6), the principal winding, N_1 and [N_2] being outside the shields, as before, with the amplifier in[put] winding, N_3, inside. The shields now ensure that $N_1i_1 - N_2i_2$ tends to zero as G is increased, so that by shielding T_1, the curr[ent] ratio i_1/i_2 may also be made precise.

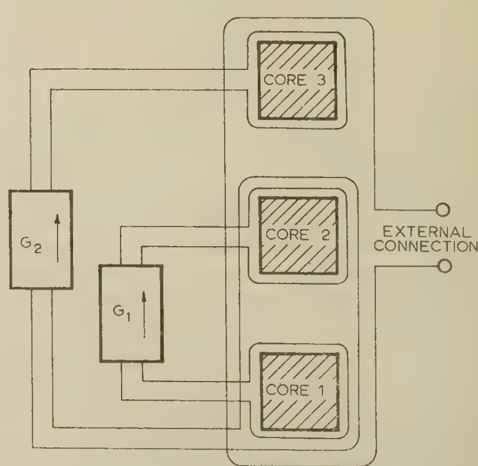

Fig. 7.—Two-stage exciting-current reduction.

ne interesting conclusion emerges that, if a composite core is e, consisting of two sub-cores, each provided with a shielded iary winding and connected by a voltage amplifier with the ropriate polarity, transformers may be wound upon it ng voltage and current ratios approximating much more ely to the turns ratio than can be achieved with the core erial alone.

(5) GENERAL

final point is that there is ideally no limit to the number mes the impedance-multiplication technique may be repeated. -stage device is shown schematically in Fig. 7. Whether it ossible to achieve a higher overall multiplication by this means re the feedback problems become insuperable has not been stigated.

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SCOTTISH ELECTRONICS AND MEASUREMENT GROUP: CHAIRMAN'S ADDRESS

By J. STEWART, M.A., B.Sc., Member.

'ELECTRONICS, THE ENGINEER AND RELIABILITY'

(Abstract *of Address delivered at* Edinburgh 11*th October, and* Glasgow 17*th October*, 1960.)

rom its earliest days electronic equipment has not enjoyed a good name for reliability, owing, no doubt, to experience early domestic radio receivers, and with some military ipment during the Second World War. Reasons for unrelia-ty were inadequate development due to a highly competitive ket in the former case and the tactical necessity for the mature release of partially developed equipment to the hting Services in the latter. After the war the expected solidation of design did not take place, because the introduc-of new ideas, materials and components continued to be too id for proper engineering to keep pace. Ultimately the stion of reliability became too serious to be ignored and the ject is now being properly studied.

ngineers are coming to realize that reliability will not occur identally, but that it must be built into the equipment. In past much equipment has been manufactured with too little ntion to proper engineering. A circuit which works in the elopment laboratory will not necessarily work on the omer's premises if it has been designed empirically and if attention has not been paid to component tolerances and hanical robustness. In particular, a great deal of trouble arisen from a failure to allow for the necessarily wide rances on valve characteristics. To a lesser degree this also lies to components. Solutions to this problem of tolerances complicated and can be costly. It is almost axiomatic to that reliability cannot be achieved without an increase in s.

he continuous introduction of new components and materials ot conducive to the establishment of designs of tried relia-y. In a competitive market it is dangerous to be out-of-date, and it requires a high degree both of competence and determination on the part of the design engineer to ensure both modernity and reliability.

It is the opinion of the author that there is too great an emphasis, in published matter on reliability, on the statistics of reliability of equipment and components, and far too little on the requirements for designing for reliability. Too many designers excuse themselves by sheltering behind the (reputed) inadequacies of the components available, instead of admitting deficiencies in engineering.

Previous strictures on Government Departments regarding the failure to pay enough attention to engineering are no longer applicable, and the current requirements for panclimatic design are most comprehensive. Developments to meet these have emphasized the fact that reliable equipment must cost more and much of the work carried out on behalf of the Government Departments would have been much too expensive in a competitive commercial field.

It is suggested that, although there is no substitute for experience, instruction in the principles of good practice in electronic design should not be left entirely to the postgraduate stage but should be begun during the college course. The subject is analogous to that of strength of materials in a mechanical engineering course. It is considered that an electronic engineer is at a grave disadvantage if he starts his professional life with insufficient knowledge of the fundamental properties of the components and devices available to him.

Electronic equipment can and will be made reliable, but only after the active application of principles of sound design and painstaking attention to detail. Reliability will not happen by itself, and the engineer who fails to build it in is failing in his duty.

Mr. Stewart is with Ferranti Ltd.

STORAGE TIME OF A TRANSISTOR WITH A DECAYING TURN-OFF CURRENT

By D. M. Taub, M.Sc., B.Sc.(Eng.), Associate Member.

(*Communication first received* 22nd *November*, 1960, *and in revised form* 9*th March*, 1961.)

One of the most common ways of using a transistor is as a switch (see Fig. 1). The transistor is switched on by applying

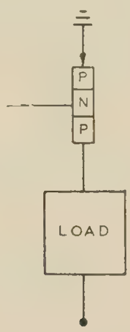

Fig. 1.—Simple transistor switch.

a forward bias to the base–emitter junction and is switched off by applying a reverse bias. It is well known that when the transistor is switched off the load current continues to flow for a time t_1 due to storage of minority carriers in the base. Expressions for t_1 have been given by Moll[1] and Beaufoy and Sparkes.[2] Using Beaufoy and Sparkes's notation,

$$t_1 = \tau_S \log_\varepsilon \frac{I_{B1} - I_{B2}}{\frac{I_C \tau_C}{\tau_B} - I_{B2}} \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where τ_S = Saturation time-constant of the transistor.
τ_C = Collector time-constant.
τ_B = Base time-constant.
I_{B1} = Base current flowing while the transistor is switched on.
I_{B2} = Base current flowing during storage time t_1.

The same sign convention is used for I_{B1} and I_{B2}, i.e. positive values represent currents causing the transistor to conduct.

In deriving eqn. (1) I_{B2} was assumed constant. Often this is not so, however; for example, in Fig. 2 an inductor is used to give I_{B2} a high initial value and so to reduce t_1. In Fig. 3 the same effect is obtained with a capacitor. With either of these circuits the expression for base current is of the form

$$i_{B2} = I_P + I_Q\varepsilon^{-t/\tau} \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

where i_{B2} = Instantaneous value of base current during t_1.
τ = Decay time-constant.

and I_P and I_Q depend on the circuit voltages and resistances.

The storage time, t_1, can be further reduced by using an inductor and capacitor as shown in Fig. 4.

Generally the component values will be chosen to produce an overdamped circuit, in which case i_{B2} is given by

$$i_{B2} = I_W + I_X\varepsilon^{-At} + I_Y\varepsilon^{-Bt} \quad . \quad . \quad . \quad . \quad (3)$$

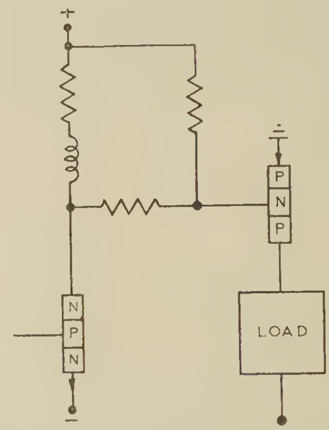

Fig. 2.—*LR* coupling circuit.

The resistor on the right provides a path for the inductive current at end of st[illegible] time.

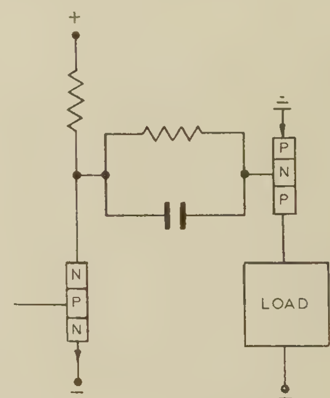

Fig. 3.—*CR* coupling circuit.

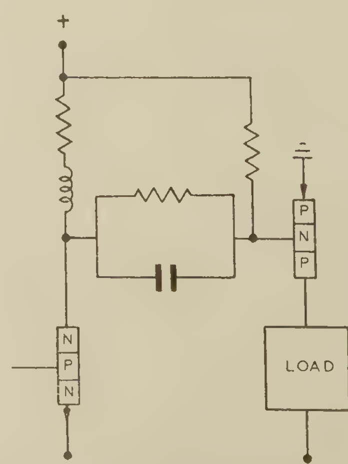

Fig. 4.—*LCR* coupling circuit.

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n the present note a general expression is derived for t_1 under ditions of varying i_{B2}. It is applied to the cases where i_{B2} s given by eqns. (2) and (3) and to the case of the under-nped *LCR* circuit. Finally two worked examples are given.

General Case

Vhile the transistor is conducting the total base charge can considered as consisting of a charge Q_B, just sufficient to intain the collector current I_C, and an excess charge Q_{BS}. e base current I_{B1} is normally designed to be just large enough ustain I_C in a transistor having minimum current gain. Since current gain varies widely between transistors of the same e this value of base current will be large enough to make $_S$ considerable in an average transistor. From eqn. (10) of ference 2 the base current I_B required to maintain the base rge Q_B is

$$I_B = \frac{Q_B}{\tau_B} = \frac{I_C\tau_C}{\tau_B} \qquad (4)$$

nis neglects the cut-off current I_{CO}.)

From eqn. (34) of Reference 2 the base current I_{BS} necessary maintain Q_{BS} is

$$I_{BS} = \frac{Q_{BS}}{\tau_S} \qquad (5)$$

lding eqns. (4) and (5),

$$I_{B1} = I_B + I_{BS} = \frac{I_C\tau_C}{\tau_B} + \frac{Q_{BS}}{\tau_S}$$

d rearranging,

$$Q_{BS} = \tau_S\left(I_{B1} - \frac{I_C\tau_C}{\tau_B}\right) \qquad (6)$$

Consider now the conditions at an instant t during the storage ne t_1. The total base charge consists of a constant part $= I_C\tau_C$ necessary to maintain I_C and an excess charge whose tantaneous value is q_{BS}. The base current can be considered maintaining these two charges and causing q_{BS} to change at ate dq_{BS}/dt. (This neglects the current required to alter the arge on the junction capacitances.) Thus,

$$i_{B2} = \frac{I_C\tau_C}{\tau_B} + \frac{q_{BS}}{\tau_S} + \frac{dq_{BS}}{dt} \qquad (7)$$

earranging and multiplying by the integrating factor ε^{t/τ_S},

$$\varepsilon^{t/\tau_S}\frac{dq_{BS}}{dt} + \frac{\varepsilon^{t/\tau_S}}{\tau_S}q_{BS} = i_{B2}\varepsilon^{t/\tau_S} - \frac{I_C\tau_C}{\tau_B}\varepsilon^{t/\tau_S}$$

$$\frac{d}{dt}(q_{BS}\varepsilon^{t/\tau_S}) = i_{B2}\varepsilon^{t/\tau_S} - \frac{I_C\tau_C}{\tau_B}\varepsilon^{t/\tau_S}$$

is is integrated with respect to t between the limits $t = 0$, ere $q_{BS} = Q_{BS}$, and $t = t_1$, where $q_{BS} = 0$, giving

$$-Q_{BS} = \int_0^{t_1} i_{B2}\varepsilon^{t/\tau_S}dt + \frac{I_C\tau_C\tau_S}{\tau_B}(1 - \varepsilon^{t_1/\tau_S})$$

bstituting for Q_{BS} from eqn. (6) and rearranging,

$$\frac{I_C\tau_C}{\tau_B}\varepsilon^{t_1/\tau_S} - \frac{1}{\tau_S}\int_0^{t_1} i_{B2}\varepsilon^{t/\tau_S}dt - I_{B1} = 0 \qquad (8)$$

LR or *CR* Circuit

The general expression for i_{B2} is given by eqn. (2). Sub-tuting this in eqn. (8) and carrying out the integration,

$$\left(\frac{I_C\tau_C}{\tau_B} - I_P\right)\varepsilon^{t_1/\tau_S} - \frac{I_Q}{1 - \frac{\tau_S}{\tau}}[\varepsilon^{(1/\tau_S - 1/\tau)t_1} - 1] - I_{B1} + I_P = 0 \qquad (9)$$

This would generally be solved by numerical or graphical methods. However, if t_1/τ_S and $(1/\tau_S - 1/\tau)t_1$ are both numerically less than unity we may make the approximation

$$\varepsilon^{nt_1} \simeq 1 + nt_1 + \frac{n^2}{2}t_1^2$$

In this way a quadratic equation is obtained and may be solved for t_1 in the normal way.

Overdamped *LCR* Circuit

Repeating the above calculation with the expression for i_{B2} given in eqn. (3) we obtain

$$\left(\frac{I_C\tau_C}{\tau_B} - I_W\right)\varepsilon^{t_1/\tau_S} - \frac{I_X}{1 - A\tau_S}[\varepsilon^{(1/\tau_S - A)t_1} - 1]$$
$$- \frac{I_Y}{1 - B\tau_S}[\varepsilon^{(1/\tau_S - B)t_1} - 1] - I_{B1} + I_W = 0 \qquad (10)$$

This may be solved in the same way as eqn. (9).

Underdamped *LCR* Circuit

In the case of an underdamped *LCR* circuit a convenient expression for i_{B2} is

$$i_{B2} = I_W + I_Z\varepsilon^{-t/\tau}\cos(ut - v) \qquad (11)$$

Care must be taken to ensure that i_{B2} remains negative throughout the decay period, otherwise the transistor will produce one or more unwanted pulses of current through the load. When one is obliged to use values of L, C and R which would allow i_{B2} to go positive, a diode should be connected across the capacitor as shown in Fig. 5. This circuit will behave as an underdamped

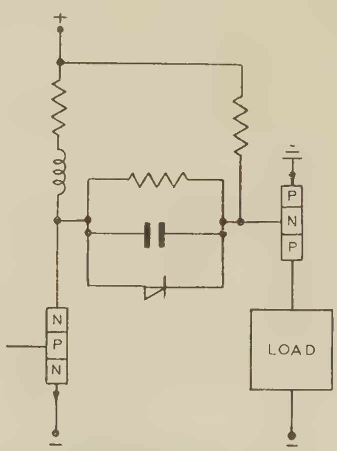


Fig. 5.—Use of diode with underdamped *LCR* circuit to prevent i_{B2} from becoming positive.

LCR circuit until the p.d. across the capacitor reaches zero unless the storage time is over before that occurs. After that time it reduces to a simple *LR* circuit. The waveforms of i_{B2} and capacitor voltage are shown in Fig. 6.

Two cases must be considered:

(i) Where no diode is necessary or where the storage time is over before it starts to conduct.

(ii) Where the diode conducts before the storage time is over.

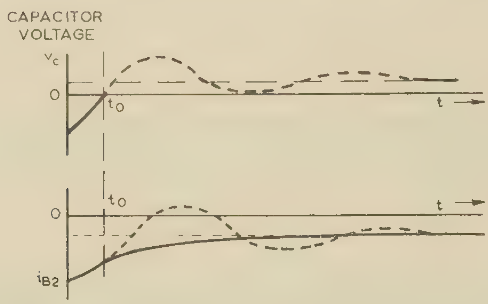


Fig. 6.—Waveforms in underdamped *LCR* circuit.
Broken lines show voltage and current in absence of diode.

Case (i) may be solved by substituting from eqn. (11) into eqn. (8) and integrating as before. This gives the following expression for t_1, which must be solved graphically or numerically:

$$\left(\frac{I_C\tau_C}{\tau_B} - I_W\right)\varepsilon^{t_1/\tau_S} - \frac{I_Z}{\tau_S\left[\left(\frac{1}{\tau_S} - \frac{1}{\tau}\right)^2 + u^2\right]}$$

$$\left\{\left[\left(\frac{1}{\tau_S} - \frac{1}{\tau}\right)\cos(ut_1 - v) + u\sin(ut_1 - v)\right]\varepsilon^{(1/\tau_S - 1/\tau)t_1}\right.$$

$$\left. - \left(\frac{1}{\tau_S} - \frac{1}{\tau}\right)\cos v + u\sin v\right\} - I_{B1} + I_W = 0 \quad . \quad (12)$$

Case (ii) requires a knowledge of the time t_0 at which the p.d. across the capacitor reaches zero. This is obtained by solving the circuit equations in the normal way. Eqn. (8) can then be rewritten

$$\frac{I_C\tau_C}{\tau_B}\varepsilon^{t_1/\tau_S} - \frac{1}{\tau_S}\int_0^{t_0} i_{B2}\varepsilon^{t/\tau_S}dt - \frac{1}{\tau_S}\int_{t_0}^{t_1} i_{B2}\varepsilon^{t/\tau_S}dt - I_{B1} = 0 \quad . \quad (13)$$

where i_{B2} is given by eqn. (11) for $0 < t < t_0$, and by eqn. (2) for $t_0 < t < t_1$. Carrying out the integrations for the general case results in a very unwieldy expression. The reader is therefore recommended to substitute numerical values early in the calculation.

Examples

Two examples will now be given, showing how the above equations are used.

Example 1.—Let us calculate the storage time of transistor VT2 in Fig. 7. It will be assumed that components, supply voltages and transistor parameters have their nominal values and that VT2 has been switched on for long enough to ensure that the stored charge in its base and the voltage V_c across the capacitor have reached their steady-state values.

When VT1 and VT2 are conducting we have

$$V_c = 12 - 0{\cdot}37 - 0{\cdot}1 = 11{\cdot}53\,\text{V} \quad . \quad . \quad . \quad (14)$$

$$I_{B1} = 11{\cdot}53/3{\cdot}9 = 1{\cdot}956\,\text{mA} \quad . \quad . \quad . \quad . \quad (15)$$

$$I_C = 11{\cdot}9/0{\cdot}24 = 49{\cdot}58\,\text{mA} \quad . \quad . \quad . \quad . \quad (16)$$

During the storage time of VT2 its base–emitter voltage will not exceed a small fraction of a volt. If this is neglected, i_{B2} will be as given by eqn. (2) where

$$I_P = -\frac{3}{3{\cdot}3 + 3{\cdot}9} = -0{\cdot}416\,\text{mA} \quad . \quad . \quad . \quad (17)$$

$$I_Q = -\left(\frac{11{\cdot}53 + 3}{3{\cdot}3} - 0{\cdot}416\right) = -3{\cdot}986\,\text{mA} \quad . \quad (18)$$

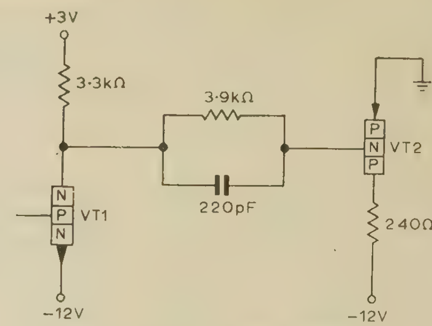


Fig. 7.—Circuit for Example 1.

Transistor characteristics:

VT1	
Collector–emitter saturation voltage	0·1 V
VT2	
Collector–emitter saturation voltage	−0·1 V
Base–emitter voltage	−0·37 V
Collector time-constant, τ_C	0·04 μs
Base time-constant, τ_B	1·4 μs
Saturation time-constant, τ_S	1·0 μs

$$\tau = 220 \times 10^{-6}\frac{3\,900 \times 3\,300}{3\,900 + 3\,300} = 0{\cdot}393\,25\,\mu\text{s} \quad .$$

Substituting the numerical values in eqn. (9), we have

$$1{\cdot}833\varepsilon^{t_1} - 2{\cdot}583\varepsilon^{-1{\cdot}5429t_1} - 0{\cdot}789 = 0 \quad . \quad .$$

The exponential terms are now expanded as shown under or *CR* circuit' giving

$$2{\cdot}157\,9t_1^2 - 5{\cdot}818\,3t_1 + 1{\cdot}539 \simeq 0 \quad . \quad . \quad .$$

whose solution is

$$t_1 \simeq 0{\cdot}297 \text{ or } 2{\cdot}399\,\mu\text{s} \quad . \quad . \quad . \quad . \quad .$$

The lower value of t_1 fulfils the conditions that t_1/τ_S a $(1/\tau_S - 1/\tau)t_1$ should be numerically less than 1. It is theref a good approximation to the solution of eqn. (20). The hig value of t_1 does not fulfil these conditions, and since eqn. (has only one real root this value may be disregarded.

A closer approximation to the solution may be made usi the Newton–Raphson method as follows. We write eqn. (in the form

$$f(t_1) = 0 \quad . \quad . \quad . \quad . \quad . \quad . \quad .$$

Then if t_{1n} is an approximate solution, a closer approximati is given by

$$t_{1(n+1)} = t_{1n} - \frac{f(t_{1n})}{f'(t_{1n})} \quad . \quad . \quad . \quad .$$

Differentiating the left-hand side of eqn. (20),

$$f'(t_1) = 1{\cdot}833\varepsilon^{t_1} + 3{\cdot}985\varepsilon^{-1{\cdot}5429t_1} \quad . \quad . \quad .$$

Now $\qquad t_{1n} = 0{\cdot}297\,\mu\text{s}$

and from eqns. (20) and (25) respectively,

$$f(t_{1n}) = 0{\cdot}046$$

$$f'(t_{1n}) = 5{\cdot}0$$

Then from eqn. (24)

$$t_{1(n+1)} = 0{\cdot}288\,\mu\text{s} \quad . \quad . \quad . \quad . \quad .$$

Closer approximations may be obtained by repeating the abc process.

:ample 2.—Let us calculate the storage time of VT2 when an :tor is added to the circuit as shown in Fig. 8. The transis- are again assumed to have been switched on for long enough he current in the inductor, the p.d. across the capacitor and stored charge in the base of VT2 to have reached their ly-state values. The p.d. between base and emitter of VT2 ain neglected during the storage time.

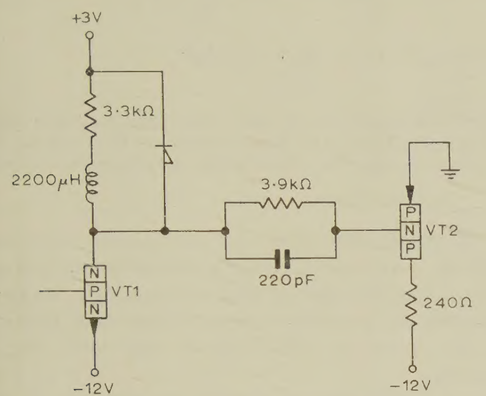


Fig. 8.—Circuit for Example 2.
Transistor characteristics as under Fig. 7.

ie circuit equations having been solved, i_{B2} is given by (11) where

$$I_W = -0{\cdot}416\,\text{mA} \quad . \quad . \quad . \quad . \quad . \quad (27)$$

$$I_Z = -5{\cdot}527\,\text{mA} \quad . \quad . \quad . \quad . \quad . \quad (28)$$

$$\tau = 0{\cdot}75\,\mu\text{s} \quad . \quad . \quad . \quad . \quad . \quad . \quad (29)$$

$$u = 1{\cdot}428\,\text{rad}/\mu\text{s} \quad . \quad . \quad . \quad . \quad . \quad (30)$$

$$v = 0{\cdot}736\,\text{rad} \quad . \quad . \quad . \quad . \quad . \quad . \quad (31)$$

$I_{B1} = 2{\cdot}956\,\text{mA}$ as in the previous example.

'e are dealing here with an underdamped *LCR* circuit and t determine whether a diode should be connected across the icitor. This is done as follows. We calculate the time t_0 at ch the capacitor voltage v_c would reach zero if t_1 were greater ι t_0. Solving the circuit equations,

$$v_c = 1{\cdot}625 - 17{\cdot}478\varepsilon^{-1{\cdot}333t}\cos(1{\cdot}428t + 0{\cdot}719) \quad . \quad (32)$$

t_0 is thus given by

$$1{\cdot}625 - 17{\cdot}478\varepsilon^{-1{\cdot}333t_0}\cos(1{\cdot}428t_0 + 0{\cdot}719) = 0 \; . \quad (33)$$

The Newton–Raphson process may be used to solve this equation, the first approximation being obtained by inspection. Taking $t_0 = 0{\cdot}5\,\mu\text{s}$ as a first approximation the second approximation is

$$t_0 = 0{\cdot}473\,\mu\text{s} \quad . \quad . \quad . \quad . \quad . \quad . \quad (34)$$

We now put t_1 equal to this value on the left-hand side of eqn. (12) and see if the result is greater or less than zero.

Substituting numerical values, eqn. (12) becomes

$$\begin{aligned} f(t_1) = {} & 1{\cdot}833\varepsilon^{t_1} + 2{\cdot}5706\varepsilon^{-0{\cdot}333t_1} \\ & [1{\cdot}428\sin(1{\cdot}428t_1 - 0{\cdot}736) \\ & - 0{\cdot}333\cos(1{\cdot}428t_1 - 0{\cdot}736)] \\ & - 0{\cdot}2734 = 0 \quad . \quad (35) \end{aligned}$$

from which $f(0{\cdot}473) = 1{\cdot}748$

Since this is greater than zero eqn. (35) could be satisfied for $t_1 = 0{\cdot}473\,\mu\text{s}$ if I_{B1} were increased, that is if the charge stored in the base of VT2 were increased. Clearly then t_1 must be less than $0{\cdot}473\,\mu\text{s}$; its value is therefore given by eqn. (35) and no diode is necessary across the capacitor.

Eqn. (35) is solved by the Newton–Raphson process using $0{\cdot}474\,\mu\text{s}$ as a first approximation. This gives after three iterations

$$t_1 = 0{\cdot}23\,\mu\text{s} \quad . \quad . \quad . \quad . \quad . \quad . \quad (36)$$

References

(1) Moll, J. L.: 'Large Signal Transient Response of Junction Transistors', *Proceedings of the Institute of Radio Engineers*, 1954, **42**, 12, p. 1773.

(2) Beaufoy, R., and Sparkes, J. J.: 'The Junction Transistor as a Charge-Controlled Device', *A.T.E. Journal*, 1957, **13**, p. 310.

Acknowledgment

The writer wishes to thank the Director of IBM British Laboratories for permission to publish this note.

THE TRAINING OF OVERSEA GRADUATE ENGINEERS

With Particular Reference to the F.B.I. Scholarships Scheme

By W. ABBOTT, C.M.G., O.B.E., Ph.D., B.Sc.(Hons.), M.I.Mech.E.

This paper, a summary of which is given below, was received 11*th June,* 1960, *published individually in October,* 1960, *and read at a joint mee of* THE INSTITUTION OF CIVIL ENGINEERS, THE INSTITUTION OF MECHANICAL ENGINEERS *and* THE INSTITUTION OF ELECTRICAL ENGIN 3*rd November,* 1960. *It was republished in April,* 1961, *in Part* A *of the* PROCEEDINGS (*page* 77). *Reprints of the paper and discussion (Abbott), price* 2s. *each* (*post free*) *will be available towards the end of June.*

The paper deals with problems arising from the training, in the engineering industry of Great Britain, of graduates from countries oversea, primarily from under-developed countries. The paper relates mainly to the Scholarships Scheme operated by the Federation of British Industries, a scheme which, though relatively small, deals with carefully selected men from a large number of countries.

The paper examines the philosophy of post-graduate practical training in the United Kingdom. It then assesses the extent of this training in industry and suggests that the total annual intake of graduates, both from home and oversea, is of the order of 5 500. The possibility of increasing this number is discussed.

The paper then considers the influence of the requirements of the major professional engineering institutions—civil, mecha and electrical—on schemes for the practical training of gradu After considering the types of engineer coming from overse paper discusses the possible inappropriateness of British trai methods for many of the visiting engineers, and sug improvements.

The paper concludes by considering the impact of gr increased demands which will be made upon Great Britai other countries of the West from the emerging nations i Commonwealth and from other developing countries.

Appendices give the balance-of-trade position with cou covered by the F.B.I. and Athlone Fellowship Schemes, a letter from the late Sir Claude Gibb relating to the traini Canadian graduates.

MONOGRAPHS PUBLISHED INDIVIDUALLY

Summaries are given below of monographs which have been published individually, price 2s. each (post free). Applications, quoting the serial numbers as well as the authors' names, and accompanied by a remittance, should be addressed to the Secretary.

A Broad-Band Waveguide Junction containing Dielectric. Monograph No. 437 E.

P. J. B. CLARRICOATS, B.Sc.(Eng.), Ph.D.

A method is described for obtaining a broadband impedance match between two joined waveguides of differing cross-section. The method involves the use of an axially mounted dielectric which partially fills the waveguide cross-section. With an appropriate choice of dielectric cross-sectional area and permittivity the electromagnetic field is mainly confined to the region of the rod. Under these conditions the waveguide cross-section may be abruptly changed without appreciable reflection. The case of two joined circular waveguides containing an axial dielectric rod is treated theoretically for H_{01}-mode propagation. A similar configuration is studied experimentally for the H_{11}-type mode of propagation. In both cases an appreciable reduction in reflection coefficient is demonstrated when the rod is present. The application of the principle to other waveguide cross-sections is briefly mentioned.

A Note on Optimum Linear Multivariable Filters. Monograph No. 439 M.

R. J. KAVANAGH, B.Sc., M.A.Sc., Ph.D.

The explicit solution for the optimum linear physically realizable multivariable filter involves the factorization of a power-spectra matrix into two matrices, one having all its poles in the left-half p-plane and the other having all its poles in the right-half p-plane. No general method of accomplishing this factorization has previously been available.

This note contributes a method of factorizing any power-sp matrix in the required manner. As a result, the explicit solutio the optimum filter is obtainable in a number of cases not previ solvable without resort to implicit methods. In the course of deve ing the factorization method it is shown that it is always pos to obtain a physically realizable multivariable system which transform any given set of signals into an equal number of incohe white-noise signals. Similarly it is shown that a physically reali multivariable shaping filter may always be found to transform of incoherent white-noise signals into an equal number of signals any desired power-spectra matrix.

An Experimental Proton Linear Accelerator using a Helix Struc Monograph No. 441 M.

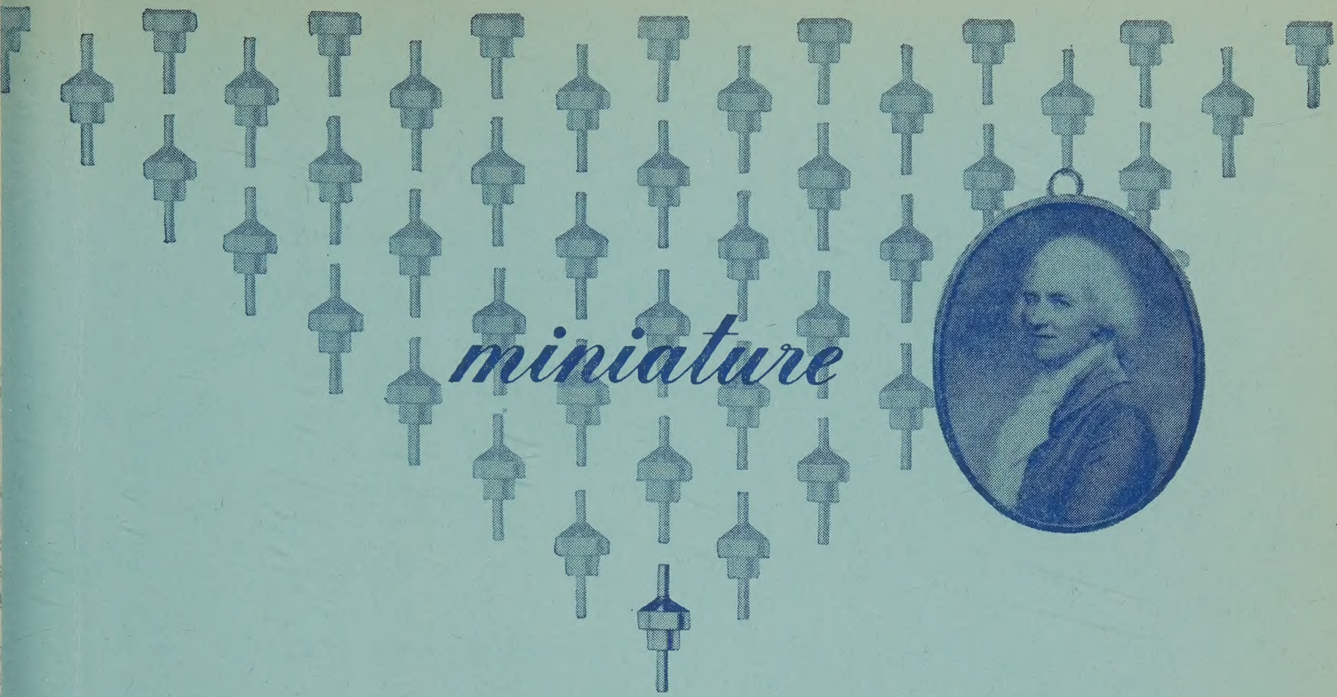
D. P. R. PETRIE, M.Sc., R. BAILEY, B.Sc., D. G. KEITH-WAL B.Sc.(Eng.), H. LONGLEY, B.Sc., and D. R. CHICK, D.Sc.

The paper describes the design and construction of a short ex mental length of linear accelerator using a helical waveguide a slow-wave structure to accelerate protons from 2·5 MeV to 4 Me

Factors influencing the design of the helix structure are consid these include the maximum voltage between turns which ca tolerated and the available power and frequency of the r.f. su The theory of a helix supported on a dielectric tube is given, and the results the variations of pitch are determined to give the req acceleration.

The accelerator was driven by a push-pull triode oscillator oper under 6μs pulsed conditions at 300 Mc/s with a peak outpu 600 kW. The oscillator and r.f. components are described in d

The beam-energy spectrum at the output of the accelerator measured for a variety of working conditions, changes being ma beam-energy input and power and frequency of the r.f. supply. results obtained confirm the theory and show that with certain re tions a helix slow-wave structure of the type described provid convenient method of proton acceleration.



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PROCEEDINGS OF THE INSTITUTION OF ELECTRICAL ENGINEERS

Part B. ELECTRONIC AND COMMUNICATION ENGINEERING (INCLUDING RADIO ENGINEERING), MAY 1961

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